Organic Electrochemical Transistor Common-Source Amplifier for Electrophysiological Measurements

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Dedicated to the memory of Professor Alasdair J. Campbell

The portability of physiological monitoring has necessitated the biocompatibility of components used in circuitry local to biological environments. A key component in processing circuitry is the linear amplifier. Amplifier circuit topologies utilize transistors, and recent advances in bioelectronics have focused on organic electrochemical transistors (OECTs). OECTs have shown the capability to transduce physiological signals at high signal-to-noise ratios. In this study high-performance interdigitated electrode OECTs are implemented in a common source linear amplifier topology. Under the constraints of OECT operation, stable circuit component parameters are found, and OECT geometries are varied to determine the best amplifier performance. An equation is formulated which approximates transistor behavior in the linear, nonlinear, and saturation regimes. This equation is used to simulate the amplifier response of the circuits with the best performing OECT geometries. The amplifier figures of merit, including distortion characterizations, are then calculated using physical and simulation measurements. Based on the figures of merit, precered electrophysiological signals from spreading depolarizations, electrocorticography, and electromyography fasciculations are inputted into an OECT linear amplifier. Using frequency filtering, the primary features of events in the bioelectric signals are resolved and amplified, demonstrating the capability of OECT amplifiers in bioelectronics.

1. Introduction

The implementation of conducting and semiconducting organic materials into classically electronic-component topologies has developed the capability to record physiological information.[11–14] An organic electrochemical transistor (OECT) consists of a source, drain, and gate electrode, where an organic polymer connects the source and drain electrodes.[15] The gate electrode is placed in an electrolyte, which itself interfaces with the organic polymer. The polymer of choice for high performance OECTs is poly(3,4-ethylenedioxythiophene):polystyrene sulfonate (PEDOT:PSS), where the PSS provides p-type doping to the conjugated PEDOT polymer.[6–8] With the injection of electrolyte ions from a gate-drain potential difference, the PEDOT:PSS dedopes, decreasing the magnitude of the current measured at the drain electrode.[9,10] OECTs function at low voltages (<1 V), are biocompatible and can be operated in aqueous environments, motivating their growing use in healthcare applications.[11–13]

There is interest in the utilization of OECTs as components in electronic circuits.[14–16] Such circuits would allow data processing local to biological environments. There are certain conditions required to implement transistors into bioelectronic circuits. They must consist of biocompatible materials, have sufficient performance parameters to measure the desired signals and must be independently gated if implemented in traditional circuit topologies.[17] Devices with a typical OECT topology that are operated in proximity would likely use a shared liquid electrolyte, not permitting individual gating. Recently, OECTs with solid-state replacements of the electrolyte have been developed and thus could be independently gated.[17,18]

Previous investigations into the use of OECTs as circuit components resulted in the development of linear and differential amplifiers.[19,20] The investigations into using an OECT as the amplifying component in a linear amplifier have primarily considered connecting the device in series with a load resistor, where the voltage gain was determined and compared for a range of load resistances.[21] However, it often is not convenient to physically vary the load resistor. An alternative method is to vary a DC bias that is applied to an AC gate voltage input and thus translating where the amplification occurs on the amplifier transfer curve. This is possible with the common-source...
amplifier circuit topology. In addition, previous investigations have not given characterizations of the linear amplifier figures of merit such as those associated with distortion effects.

In this work, the implementation of OECTs into common-source amplifier circuits was investigated. Using interdigitated-electrode OECT characteristics, conditions on the voltages and resistances used in the common-source amplifier circuit were identified. It was found that the OECTs had the capacity to be utilized in the circuit. The linear amplifier circuit was then characterized with simulations and physical measurements with an emphasis on determining the performance parameters described by the figures of merit of total-harmonic and intermodulation distortions, gain, input-referred noise, and bandwidth. The circuit was simulated through the formulation of an equation for transistor operation which showed a good fit to transistor characteristics across the linear, nonlinear, and saturation regimes. Prerecorded electrophysiological signals from spreading depolarizations (SDs), electrocorticography (ECoG), and electromyography (EMG) fasciculations were inputted at the gate electrode and amplified using the circuit.

2. Results and Discussion

2.1. Transistor Fabrication and Characterization

Organic electrochemical transistors were fabricated with interdigitated electrodes, where there were two interdigitated styles, each with two sizes. The styles used were those of comb and spiral geometries. The comb electrode devices had approximately square active layers with a side length of 82 µm (small) or 174 µm (large). The spiral electrode devices had approximately circular active layers with a diameter of 81 µm (small) or 182 µm (large). The devices were constructed with interconnects of an array geometry. The devices were further divided into those with an interconnect thickness of 200 nm and an active layer of 70 nm or those similarly with thicknesses of 300 and 35 nm, respectively. In total there were eight variations of OECT geometry considered in this investigation. The exact geometries of these transistors are as described in previous investigations.[22] Optical micrographs of the OECT source and drain electrode, encapsulation, and active layers are provided in Figure S1, Supporting Information. The electrodes were composed of gold with a chromium adhesive layer and were fabricated on a 125 µm polyethylene terephthalate (PET) substrate. An encapsulation layer of SU-8 was coated with a thickness of 1 µm. The active layer consisted of the organic conducting polymer PEDOT:PSS. Each layer was patterned using photolithography, allowing for µm-scale device features. The fabrication method was that as described in previous research.[22] For each device, the output, transfer, and response time characteristics were found.

2.2. Linear Amplifier Circuit

The OECT common-source linear amplifier circuit topology is shown in Figure 1. The amplifier is powered by a supply voltage, $V_{DD}$, which is used to amplify an input voltage, $V_{IN}$. The potential divider circuit formed with $V_{DD}$, $R_1$, and $R_2$ then provides a tunable DC component to $V_{IN}$, resulting in the input gate-ground voltage to the OECT, $V_G$. This allows the selection of the region of amplification within the OECT and thus the output gain and linearity. The supply voltage also provides a potential difference to the drain resistor, $R_D$, the OECT and the source resistor, $R_S$, in parallel with the bypass capacitor, $C_1$. The amplifier gain and the range of amplifiable gate voltage inputs are dependent on the values of the source and drain resistors and the supply voltage. The bypass capacitor acts to suppress high-frequency signals that are present in the AC voltage component of the potential difference across the source resistor, drain resistor, and OECT.

In addition to the voltage divider, there is typically a contribution to the gate-ground voltage at the OECT gate from an input capacitor, which removes the DC component of $V_{IN}$ and allows the DC gate component to be controlled entirely through varying resistors $R_1$ and $R_2$. This capacitor also acts as a high-pass filter to AC voltages originating from $V_{IN}$. However, this investigation sought to build and fully characterize a DC-coupled amplifier which could amplify a wide range of frequencies characteristic of bioelectric inputs. As such, the attenuation of low-frequency voltages from the input capacitor was not desired. Therefore, the use of an input capacitor, and thus the removal of low-frequency voltages, was avoided by using a waveform generator to produce $V_{IN}$, where its output voltage was equivalent to that otherwise produced if the capacitor was present. As a result, the amplifying system developed in this work was completely DC-coupled, allowing the study of its behavior in a wide frequency range (from DC up to the 10 kHz-scale).

The drain current of the OECT is determined by the loadline of the amplifier, which describes the dependence of drain current on the circuit topology. In this investigation, the bypass capacitor value, source resistor value, and the frequencies inputted in the amplifier characterization and the bioelectric signal amplification allowed the total drain current, which is the summation of AC and DC components, to be approximately

![Figure 1. Common source linear amplifier circuit using input voltage $V_{IN}$, potential divider resistors, $R_1$ and $R_2$, source resistor, $R_S$, drain resistor, $R_D$, bypass capacitor, $C_1$, and supply voltage $V_{DD}$. The OECT is represented by the three-terminal device with gate, source, and drain labeled $G$, $S$, and $D$, respectively, through which a total drain current, $I_D$, flows.](image-url)
described by only the DC load line (see Supporting Information). Therefore, consideration of only the DC load-line provided approximate insights into amplifier behavior for low-frequency AC voltage inputs. As such, application of only the DC load line is considered in the following analyses. Furthermore, the total drain current in this approximation, which is the summation of DC and low-frequency AC components, is subsequently denoted as $I_D$.

Gate voltage relations in the circuit topology presented in Figure 1 are described by

$$V_{GS} = V_C - V_S = \frac{R_2}{R_1 + R_2} V_{DD} - I_D (V_{GS}, V_{DS}) R_S$$  

where $V_{GS}$ is gate-source voltage, $V_{DS}$ is drain-source voltage, $V_C$ is gate-ground voltage, $V_S$ is source-ground voltage and all other parameters have been previously described. The DC load-line, as reduced from Equation (S1), Supporting Information, describes the distribution of supply voltage across the OECT, source resistor, and drain resistor:

$$V_{DD} = V_{DS} + I_D R_S + I_D R_D$$  

As shown in Equation (1), the gate-source potential difference is dependent on the drain current of the OECT, which itself is dependent on gate-source voltage. As the negative gate-ground voltage increases, the drain current increases and thus the potential difference across the source resistor also increases (both decrease in magnitude toward zero). Thus, an increase in $V_C$ leads to an increase in $V_S$, which stabilizes $V_{GS}$. As the gate-ground voltage increases, the potential difference across the drain resistor also increases and the drain-source voltage decreases. However, if the device is operating in the saturation regime then changes in drain-source voltage will not induce any further changes in drain current than those already caused by gate voltage variations. As such, the device should be operated in the saturation regime for amplification. Furthermore, short-channel effects were not desirable and were not observed with 4 µm channel lengths on a subset of the OECTs.

The overlap of the load-line with the saturation region of the output characteristics shows the range of gate-source voltages where the device can be operated. As shown by the output characteristics, positive gate-source voltages are required to operate the device in this regime, where $[V_{GS} - V_{DS}] > V_{th}$ for OECT pinch-off voltage $V_D$. A condition set on the difference between the applied gate-source and drain-source voltages is the hydrolysis level, which occurs at ~1.2 V. Operation at greater potential differences can cause device degradation and thus only voltage combinations below this limit are allowed. As the value of the supply voltage, $V_{DD}$, is required to be negative to operate the OECT in the saturation regime, the range of applied voltages is further restricted. This is because the voltage divider circuit can only apply a negative DC bias to the input AC signal. For negative gate-ground voltages to correspond to positive gate-source voltages, it can be seen from Equation (1) that, in approximating low frequency AC signals with DC behavior, the condition $V_S \equiv I_D R_S < V_C$ is required. With negative current values in the saturation regime, the source resistor thus acts to translate negative gate-ground inputs into positive gate-source potentials.

The output voltage, $V_{OUT}$, was defined to be the drain-source voltage across the OECT, $V_{OUT} \equiv V_{DS}$. The drain-ground voltage was also considered; however, $V_{DS}$ decreases as $V_S \equiv I_D R_S$ increases, as can be seen on the load-line output characteristics. Therefore, the opposing voltages result in smaller changes, and thus a smaller gain, relative to considering just $V_{DS}$. The gain was defined as the fraction of output voltage range to gate-ground voltage input range, $\Delta V_{OUT}/\Delta V_{G}$. The gain is dependent on the summation of the source and drain resistors and has representation in the gradient of the load-line. As the gradient decreases in magnitude, there is a greater change in $V_{DS}$ for a given change in $V_{GS}$. However, given the hydrolysis limit, the range of gate-source voltages that the load-line intersects in the saturation region decreases and thus the input range decreases. Therefore, there is a trade-off between gate input range and gain.

For clinical application, the amplifier could be implemented within an electrophysiological recording device as a component that measures and amplifies bioelectric activity. Given that the amplifier is completely DC-coupled, as it was in this investigation, it could potentially be used in ECoG applications where the device directly interfaces the cortical surface. The cerebrospinal fluid would act as the OECT electrolyte and the gate voltage produced by the neuronal ionic fluxes would be shifted by the DC-bias produced by the potential divider (see Supporting Information). It has been previously shown that when applying a constant low voltage for the operation of an OECT to measure activity in a neuronal volume, there has been no measurable evoked neuronal activity or disturbance of the extracellular environment. The neuronal activity would then be amplified by the circuit, producing an output voltage.

In addition to measuring and amplifying bioelectric activity, the circuit could be used to amplify a bioelectric signal measured from an electrode. The electrode’s measured signal would be the $V_{IN}$ and the low-voltage amplifier could be utilized local to the physiological environment due to its biocompatible properties. The circuit would need to use an electrolyte that is physically separated from the physiological environment to avoid the amplification of signals from other spatial locations. Alternatively, a solid-state replacement to the electrolyte could be used. Such a replacement would also be needed if using multiple amplifiers taking electrode signals from a number of recording sites as each OECT would require independent gating. Finally, the device could be used for providing a voltage output which describes the concentration of chemical species. The OECT active layer or gate electrode could be functionalized to respond to the presence of a particular species in the electrolyte. This could thus be applied to measuring and amplifying the changes in chemical concentrations of the different ionic species present in physiological electrolytes such as cerebrospinal fluid.

### 2.3. DC Load Line Analysis

Using the OECT output characteristics from this and the previous work, values of $V_{DD}$, $R_S$, and $R_D$ were iterated to find combinations that satisfied all OECT linear amplifier criteria. The iteration varied the supply voltage between −3.0 to −1.0 V
in 0.1 V steps and the source and drain resistors between 1 to 750 Ω in 10 Ω steps. Each combination (of 112500) resulted in a DC load-line passing through the output characteristics, and the analysis was applied to a number of transistors of each OECT geometry. For combinations that satisfied the criteria, the values of the range of input gate-ground voltages and the corresponding changes in drain-source voltage were found. For each OECT, the values of $V_{DD}$, $R_S$, and $R_D$ which maximized the gain were identified and the associated gate-ground range was found. As such, for OECTs of a given geometry, the mean and standard deviation values were found for the number of successful parameter combinations, maximum gain values, and the associated gate-ground ranges. Results of this DC load-line analysis are shown in Table 1 alongside the mean and standard deviation of the pinch-off voltage for each OECT geometry. Although the analysis was restricted by the voltage range and resolution of the output characteristics, it was still indicative of the suitability of the different OECT types to be applied to the linear amplifier circuit.

It was found that the 35 nm devices provided greater gain values in the DC load-line analysis. The 70 nm devices had a greater gate-ground input range; however, the gate input range for the 35 nm devices was still significantly greater than the limits of extracellular bioelectric voltages. As the pinch-off voltage decreased, which was observed in the 35 nm devices relative to the 70 nm devices, the gain increased (see Figure S3, Supporting Information). For a given load-line with defined gradient and fixed gain, the gate-source voltage range would increase if the device entered the saturation region at a lower magnitude of drain-source voltage. Similarly, for a fixed gate-source voltage range, the drain-source voltage range, and thus gain, would increase given saturation at a lower magnitude of drain-source voltage. Such performance improvements require lower values of pinch-off voltage, which Bernards et al., showed was proportional to the thickness of the organic active layer.[9] The pinch-off voltage can also be modulated through varying other OECT parameters (see Supporting Information). The channel thickness dependence suggests that the gain of an OECT common-source linear amplifier can be increased using a thinner charge transport layer. It has also been shown that the OECT response time, which determines the amplifier bandwidth, is proportional to active layer thickness.[9] Therefore, in decreasing the thickness, it is further suggested that the gain and bandwidth can be simultaneously increased under this amplifier configuration. These insights suggest that a common-source amplifier topology, as implemented for electrophysiology using OECTs in this configuration, would have improved performance parameters with a thinner OECT active layer.

It was found that the gate-source voltage range that frequently satisfied the conditions were $0.1 \, V < V_{GS} < 0.2 \, V$. In this range, the transfer characteristics showed that there was an approximately linear relationship between drain current and gate voltage. This allowed a further analysis to determine the gain values if the gate-source voltage range was reduced from 100 mV to the mV-scale, as suitable for a range of bioelectric signals.

The relationship between the original gate-source voltage range $\delta$ with gain $G$ and the new range $\gamma = \gamma \delta$, $\gamma > 0$ with gain of $G'$, can be analytically determined when assuming a linear current–voltage relationship in the range $\delta$. It can be shown that

$$G' = (\gamma - 1)/\gamma + G/\gamma$$  \hspace{1cm} (3)

Furthermore, the fraction in gate-source voltage range is equivalent to the fraction in gate-ground voltage given that only the drain resistor is increased to reduce the load line gradient (the source resistor is left constant) and the OECT transfer curve is linear throughout the original gate-source voltage range $\delta$ (see Supporting Information). In this investigation, the output characteristics gave a gate-source range of $\delta = 100 \, mV$ with a gain value of $-G = -2$. For a gate-source voltage range of 2 mV, then $\gamma = \delta/\delta = 0.02$, predicting a gain of $G' = -149$. In limiting the gate-source voltage range, the load line gradient decreases in magnitude. As such, the source and/or drain resistance values must increase. In turn, the $V_{DD}$ must increase to maintain an adequate drain-source voltage to keep the device in the saturation regime.

2.4. Simulation of Device Behavior

An equation was formulated which provides a good fit to transistor output characteristics across the linear, nonlinear,
and saturation regimes. This equation was developed and utilized because it provided continuity across the three regimes and thus eliminated any discontinuity that would occur from switching between equations at regime boundaries. Furthermore, an equation of this form had not been identified in previous literature and so the capabilities of this novel equation were investigated.

To derive the equation, the properties of the output characteristics were considered. First, the equation needed to pass through the origin and have a linear region which tended to an asymptote with a monotonic intermediary region. As such, the product of a linear variable of zero intercept with a sigmoid function was considered. It was found that the translated arc tangent function provided suitable properties. Therefore, the product was parameterized with \( \rho \) and \( \lambda \) as

\[
f(x) = \rho x \left[ \arctan(\lambda x) - \frac{\pi}{2} \right]
\]

(4)

Second, to translate Equation (4) from \( y = f(x) \) to \( y' = f(x - \mu) + \nu \) whilst maintaining \( y' = x = 0 \), it is clear that \( \nu \) is constrained as \( \nu = -f(-\mu) \), and thus \( y' = f(x - \mu) - f(-\mu) \). Therefore,

\[
y' = f(x - \mu) - f(-\mu) = \rho(x - \mu) \left[ \arctan\left(\lambda(x - \mu)\right) - \frac{\pi}{2} \right] + \rho \mu \left[ \arctan(-\lambda \mu) - \frac{\pi}{2} \right]
\]

(5)

Finally, the variables and parameters in Equation (5) were associated with OECT characteristics. The most appropriate relations for p-type depletion mode characteristics found that \( y' = I_D, x = V_{DS}, \rho = -G/3, \) and \( \mu = \alpha V_{CS} + \beta V_F \) with \( \alpha > 0 \) and \( \lambda < 0 \). Therefore, the equation states that the drain current, \( I_D \), varies as

\[
I_D = -\frac{G}{3} \left[ V_{DS} - (\alpha V_{CS} + \beta V_F) \left\{ \arctan\left(\lambda \left(V_{DS} - (\alpha V_{CS} + \beta V_F)\right) - \frac{\pi}{2}\right) \right\} + (\alpha V_{CS} + \beta V_F) \left\{ \arctan\left(-\lambda \left(\alpha V_{CS} + \beta V_F\right)\right) - \frac{\pi}{2}\right\} \right]
\]

(6)

where \( G \) is the OECT conductance, \( V_{DS} \) is the drain-source voltage, \( V_{CS} \) is the gate-source voltage, \( V_F \) is the pinch-off voltage, \( \alpha \) and \( \beta \) are dimensionless parameters and \( \lambda \) is a parameter with units of inverse voltage, and \( \alpha, \beta, \) and \( \lambda \) are identified through applying a best-fit to the output characteristics. Using Equation (6), the transconductance of a modeled transistor at a constant drain voltage is given as

\[
g = \frac{\partial I_D}{\partial V_{CS}} = \frac{G}{3} \left[ \arctan\left(\lambda \left(V_{DS} - (\alpha V_{CS} + \beta V_F)\right)\right) - \arctan\left(-\lambda \left(\alpha V_{CS} + \beta V_F\right)\right) \right] + \frac{V_{DS} - (\alpha V_{CS} + \beta V_F)}{\lambda \left(V_{DS} - (\alpha V_{CS} + \beta V_F)\right) + 1/\lambda} + \frac{\alpha V_{CS} + \beta V_F}{\lambda \left(V_{DS} - (\alpha V_{CS} + \beta V_F)\right) + 1/\lambda}
\]

(7)

Equation (6) can also be used to model p-type enhancement mode devices. Furthermore, n-type enhancement and depletion mode devices can be modeled using \( \rho = G/3, V_{DS} \rightarrow -V_{DS}, \alpha < 0, \) and \( \lambda < 0 \) (see Figure S5, Supporting Information). The equation can be used as a tool for simulating the approximate dynamics of transistors under varying inputs, such as the changes in drain current under simultaneous changes in drain-source and gate-source voltages. Furthermore, circuits involving the use of transistors can be approximately simulated.

Temporal dynamics were also introduced by using the analyses from Bernards et al., which derived an exponential response in drain current. When assuming electronic transport time is much smaller than ionic transport time, the drain current can be described by

\[
I_D(t + \delta t, V_{CS}) = I_{D,SS}(t, V_{CS}) + \Delta I_{D,SS}(V_{CS} - V_{GD})e^{-\frac{t}{\tau}}
\]

where \( I_{D,SS} \) is the steady-state drain current and \( \tau \) is a response time constant which gives a 90% response time of \( \tau \). The output, transfer, and transconductance-frequency characteristics as described by Equation (6) were shown in Figure 2a,b,d, respectively, with the parameters of \( G = 10 \times 10^{-3} \), \( V_F = 0.6 \), \( \tau = 43 \mu s \) (giving a 90% response time of 100 \( \mu s \) \( \alpha = 1.2, \beta = -0.6, \) and \( \lambda = -3.5 \)). A fitting of Equation (6) was applied to the OECT output characteristics of an OECT with the 35 nm small spiral geometry, as presented in Figure 2c. It was found that the OECT characteristics were reproduced with an accuracy sufficient for further utilization. Varying the \( \alpha, \beta, \) and \( \lambda \) parameters in Equation (6) gave physically interpretable effects including those related to the intrinsic properties of the organic active layer (see Supporting Information).

With Equation (6) showing capability to reproduce the features of the output characteristics, an approximate amplifier transfer curve was determined. With the circuit parameters which provided maximum gain as found in the DC load-line analysis, the gate-ground voltage input and corresponding drain-source output voltages were calculated. Analogous to the transconductance, the gain was found by taking the derivative of the amplifier transfer curve. The gate-ground voltage is of primary interest in such a transfer curve, though the equivalent gate-source voltage relationship can also be found. The transfer curve was found by computationally solving Equation (2) with the \( I_D \) given in Equation (6), giving the allowed drain-source and gate-source voltage combinations (as shown in Equation (S10), Supporting Information). The equivalent gate-ground voltages were then found by using Equation (1). The relationship between gate-source voltage and gate-ground voltage is approximately linear (see Figure S7, Supporting Information). Such transfer curves are shown for the 35 nm large comb and 35 nm small spiral OECTs in Figure 3a,c, respectively.
The temporal response of the amplifier was thought to be limited only by that of the OECT. Inspired by Bernards et al., to simulate the approximate temporal response of the linear amplifier, the output voltage was defined as

\[
V_{\text{OUT}} = V_{\text{G}} + \Delta V_{\text{OUT}}(t) = V_{\text{G}} + \frac{V_{\text{DS}} - V_{\text{G}}}{g_{\text{m}}}e^{-t/\tau},
\]

where \( V_{\text{OUT}} \) is the DC output voltage across the OECT, \( V_{\text{G}} \) is the gate-ground voltage and \( \tau \) is the OECT response time constant.\(^9\) This provided a comparison between the noise-free approximation and the physical measurement of the amplifier characterization. To demonstrate the effects of varying resistor \( R_2 \) on the amplifier circuit, the output voltage was simulated with a 1 Hz, 20 mV amplitude sinusoidal gate input. The response of this simulation is shown in Figure 3b,d, which illustrated that as the gate-ground voltage increased from a decrease in the value of \( R_2 \), the gain of the amplifier increased.

As designed in determining the voltages and resistances in the DC load-line analysis, the gate-drain voltages within hydrolysis were associated with negative gate-ground voltages. Furthermore, the peak gain values were within the ranges expected in Table 1, supporting the capabilities of Equation 6. The amplifier transfer curves in Figure 3a,c highlight that the region of gate-ground voltage used can significantly affect the amplification. Within the hydrolysis limit, the gain varies greatly in the gate-ground voltage range, emphasizing the importance of identifying the optimal operating point to maximize the linearity of the amplified output. The topology of the common-source linear amplifier allows for the region of amplification to be chosen through the variation of the potential divider resistors \( R_1 \) and \( R_2 \), as illustrated in Figure 3b,d. The results in Figure 3a,c suggest that, for each circuit topology in this investigation, the optimal operating point existed around the point of maximum gain.

2.5. Linear Amplifier Figures of Merit

Using the optimal circuit parameters found with the DC load-line analysis, five OECTs each from two geometry types were physically tested in the linear amplifier circuit. The device types chosen were the 35 nm small spiral and 35 nm large comb. These were used because the 35 nm devices showed a greater capability in implementation and allowed a comparison of performance with varying electrode geometry. The common-source linear amplifier circuit was physically characterized to give its figures of merit of gain, total harmonic distortion (THD), intermodulation distortion (IMD), input-referred noise, and bandwidth (see Supporting Information). The characterizations were performed using inputs within electrophysiological limits. The figures of merit were determined for both device types and a mean and standard deviation was found for each performance.

![Figure 2](image-url)
The device gain could have been increased with a decrease in gate voltage range. However, to ensure stability of the OECTs throughout testing, the chosen parameter values were those found to satisfy the OECT linear amplifier criteria in the DC load-line analysis.

In the strategy of monitoring electrophysiological signals for diagnostic and therapeutic purposes, the design and realization of a recording device that can accurately record weak signals in the presence of strong interference originating from noise sources within the clinical environment is one of the most challenging issues. Sources of such noise include the 50 Hz interference, high-frequency interference from radio frequencies, and the operating frequencies of the medical devices used in the clinic. Crucially, the picking up of one or more interference signals by an amplifying device which is characterized by narrow passbands. The IMD was found by inputting two sinusoidal waveforms with frequencies $f_1$ and $f_2$ and of equal amplitude at $V_{IN}$. The sinusoidal waveforms were increased through the same range as those in the THD measurements. The fundamental frequencies and the $2f_{1/2} - f_{1/1}$ components were used to generate the IMD third-order intercept point, IP3.

To measure the input-referred noise, the gate electrode was connected to the ground and the output voltage was divided by the gain. A periodogram was then applied to find the spectral density of the resulting signal. The power spectral density was then numerically integrated within specified frequency ranges.

**Figure 3.** a) Simulated amplifier transfer curve using an OECT of 35 nm large comb geometry with circuit parameters of $R_1 = 1 \, \text{k} \, \Omega$, $R_2 = 40 \, \Omega$, $R_3 = 90 \, \Omega$, $R_0 = 740 \, \Omega$, $V_{DD} = -2.2 \, \text{V}$, and $C_1 = 100 \, \text{nF}$ (where the $R_0$, $R_2$, and $V_{DD}$ were determined using the DC load-line analysis), transistor parameters of $G = 6.14 \, \text{mS}$, $V_T = 0.45 \, \text{V}$, and $\tau = 159 \, \mu\text{s}$ and Equation (6) parameters of $\alpha = 1.02$, $\beta = -0.62$, and $\lambda = -4.42 \, \text{V}^{-1}$. b) Simulation of amplifier output for varying $R_2$ values using the amplifier transfer curve shown in (a) using a 1 Hz, 20 mV amplitude sinusoidal gate input. c) Simulated amplifier transfer curve using an OECT of 35 nm small spiral geometry with circuit parameters of $R_1 = 1 \, \text{k} \, \Omega$, $R_2 = 55 \, \Omega$, $R_3 = 170 \, \Omega$, $R_0 = 680 \, \Omega$, $V_{DD} = -2 \, \text{V}$, and $C_1 = 100 \, \text{nF}$, transistor parameters of $G = 6.86 \, \text{mS}$, $V_T = 0.51 \, \text{V}$, and $\tau = 37 \, \mu\text{s}$ and Equation (6) parameters of $\alpha = 1.08$, $\beta = -0.64$, and $\lambda = -3.48 \, \text{V}^{-1}$. d) Simulation of amplifier output for varying $R_2$ values using the amplifier transfer curve shown in (c) using a 1 Hz, 20 mV amplitude sinusoidal gate input.

The harmonic distortion produced by single tone sinusoidal signals of various frequencies, as performed in THD, does not convey all the information required to assess in full the amplifier’s linearity performance in a clinical setting where multi-frequency noise aggressors exist. Hence, it is often required that an amplifier be assessed in terms of the IMD product levels produced by two tones positioned spectrally close to each other and applied simultaneously at the input of the amplifier. Such IMD tests allow for the practical and meaningful assessment of linearity even in the case of designs characterized by narrow passbands. The IMD was found by inputting two sinusoidal waveforms with frequencies $f_1$ and $f_2$ and of equal amplitude at $V_{IN}$. The sinusoidal waveforms were increased through the same range as those in the THD measurements. The fundamental frequencies and the $2f_{1/2} - f_{1/1}$ components were used to generate the IMD third-order intercept point, IP3.
bands to find the RMS voltage noise. The Bode plot was determined by inputting a sinusoidal waveform into the gate with varying frequency and determining the amplitude of the response. The device was swept from 1 Hz to the 10 kHz-scale and the frequency at which the amplitude changed by −3 dB defined the upper limit of the bandwidth. The physical characterization of the common-source linear amplifier using the small 35 nm small spiral OECT is shown in Figure 4, where the locations of the figures of merit are highlighted. The THD, IMD, and Bode amplifier characterizations were also simulated. As the physical characterizations only considered gate inputs within electrophysiological levels, the simulation provided the distortion levels at higher input values. Table 2 gives the mean and standard deviation values of the physically measured and simulated figures of merit for the two transistor geometries tested.

The corner frequency, defined as the noise level at which flicker and thermal noise are of the same magnitude, could not be identified in the input-referred noise because there was no observed transition between a frequency dependent and frequency independent region. A further property which could not be determined using only the physical THD characterization was the input dynamic range. Only the lower limit of the range was identifiable as the THD remained lower than 1% for all subsequent voltage inputs of greater amplitude within the electrophysiological limit. At the input amplitudes of greatest magnitude, THD calculations showed the powers of higher harmonics were increasing but a level lower than 1% was maintained. As such, Table 2 shows only the lower 1% THD limit for physical measurements. The simulation was used to investigate the nonlinearity at voltage input amplitudes greater than those used in the physical characterizations and the upper limit at which the THD returned to 1% was identified (see Figure S8, Supporting Information). With physical measurements providing a lower 1% THD value and simulation measurements providing an upper limit, an approximate input dynamic range could be calculated.

For the physical measurements of the IP3, the input powers, as applied within the bioelectric voltage limits, did not result in a clear third-order linear component. The points used for extrapolation were the highest points of the 2f₁ − f₂ component where an increase in input power corresponded to an increase in output power, as illustrated in Figure 4d. However, the gradient found in such a linear extrapolation did not take the expected value of 3. Due to the limited number of points which did show consistent increases, the inter-amplifier variation in the line of best fit was large and thus the standard deviation in the IP3 was high. The simulation of the IMD allowed greater input powers than the bioelectric limit and showed the expected behavior with less variation, as illustrated in Figure S8d,f, Supporting Information. As such, the simulated IP3 points are also given in Table 2.
Table 2. Physical and simulation measurements of the figures of merit of the common source linear amplifier with a t-test of independence of each characteristic for the small spiral 35 nm and large comb 35 nm OECT geometries.

| Property                        | Small spiral 35 nm | Large comb 35 nm | Units                  | Independent? |
|--------------------------------|--------------------|------------------|------------------------|--------------|
| Bypass capacitor, $C_t$         | 100                | 100              | nF (3 s.f.)            | N/A          |
| Resistor $R_1$                  | 1.0                | 1.0              | kΩ (2 s.f.)            | N/A          |
| Resistor $R_2$                  | 55                 | 40               | Ω (2 s.f.)             | N/A          |
| Source resistor $R_o$           | $70 \pm 6$         | $78 \pm 12$      | Ω (3 s.f.)             | N/A          |
| Drain resistor $R_o$            | $610 \pm 37$       | $698 \pm 66$     | Ω (3 s.f.)             | N/A          |
| Supply voltage $V_{DD}$         | $-1.90 \pm 0.04$   | $-2.20 \pm 0.04$ | V (2 d.p.)             | N/A          |
| Cut-off frequency               | $5090 \pm 825$     | $1000 \pm 49$    | Hz (4 s.f.)            | Yes          |
| Cut-off frequency sim.          | $4500 \pm 189$     | $969 \pm 2$      | Hz (3 s.f.)            | Yes          |
| Lower 1% THD gate phys. (peak amplitude) | $0.58 \pm 0.29$ | $0.45 \pm 0.07$ | mV (2 d.p.)            | No           |
| Upper 1% THD gate sim. (peak amplitude) | $95.68 \pm 17.50$ | $48.88 \pm 15.78$ | mV (2 d.p.)            | Yes          |
| Input dynamic range (peak amplitude) | $95.10 \pm 17.54$ | $48.43 \pm 15.76$ | mV (2 d.p.)            | Yes          |
| OECT pinch-off voltage          | $519 \pm 7$        | $453 \pm 3$      | mV (3 s.f.)            | Yes          |
| Measured [gain]                 | $1.97 \pm 0.11$    | $2.82 \pm 0.21$  | (3 s.f.)               | Yes          |
| Simulated [gain] using Equation (6) | $2.17 \pm 0.15$     | $3.22 \pm 0.12$  | (3 s.f.)               | Yes          |
| Integrated noise                |                    |                  |                        |              |
| 0.5–40 Hz                       | $87.5 \pm 37.9$    | $88.5 \pm 13.0$  | μVRMS (1 d.p.)         | No           |
| 0.5–100 Hz                      | $95.7 \pm 39.2$    | $97.2 \pm 13.4$  | μVRMS (1 d.p.)         | No           |
| 0.5–500 Hz                      | $106.0 \pm 42.4$   | $108.8 \pm 13.8$ | μVRMS (1 d.p.)         | No           |
| 0.5–5000 Hz                     | $119.9 \pm 47.7$   | $121.3 \pm 15.7$ | μVRMS (1 d.p.)         | No           |
| IP3 input power                 | $84.57 \pm 44.93$  | $60.62 \pm 39.82$| dBV (2 d.p.)           | No           |
| IP3 input power sim.            | $4.07 \pm 0.44$    | $2.68 \pm 0.31$  | dBV (2 d.p.)           | Yes          |

For each property associated with the figures of merit, a t-test of independence was performed comparing the two geometries each implemented into five amplifiers. The test assumed that the variances of the underlying populations were unequal and the p-value for significance was 0.05. The results of the statistical tests are shown in Table 2.

As found previously through the DC load-line analysis, the lower pinch-off values were associated with higher gain values presented in Table 2. Relative to the DC load-line gain in Table 1 and the simulated gain in Table 2, the physically measured gain was of the same scale and it followed that the large comb had a greater gain than that of the small spiral. It was found by Zhang et al., that the properties of active layers in OECTs fabricated using similar PEDOT:PSS formulations changed after immersion in water.[7] Therefore, the difference between the measured and simulated gain could have arisen from a change in the transistor output characteristics as a result of the immersion of the PEDOT:PSS active layer in an aqueous environment for ~1 h during the amplifier characterization.

In addition to ensuring device operation below the hydrolysis limit, a conservatively low gain was utilized because, in general, there is a trade-off between gain value and linearity levels. When the gain value increases, the output signal level is increased and the nonlinearities innate to the device are exercised strongly, leading to a subsequent increase in the distortion products. On the contrary, when the gain decreases, the output signal level is decreased and the nonlinearities innate to the device are exercised weakly, thus the distortion products are buried below the system noise floor. Furthermore, the primary aim of this study was to maximize the achieved gain of the OECT-based amplifier whilst maintaining an approximately linear relationship between drain current and gate voltage. Indeed, the measured THD and IMD levels reported in this work indicate that the aim of providing a linear amplifying system built with high-performance OECTs has been accomplished. However, in addition to varying the gain of the implemented circuit by changing only the potential divider resistor values $R_1$ and $R_2$, the gain can be increased by varying the supply voltage and the source and drain resistors, as described in Equation (3).

The results in Table 2 suggest that an increased gain was associated with a lower bandwidth. However, as previously mentioned, higher gain values and bandwidths are still thought to be associated with a thinner OECT active layer for a constant electrode geometry. The bandwidth of the amplifier was limited by the response time of the OECT, and so was dependent on the OECT gate-channel capacitance. This capacitance is reduced with a thinner active layer, though there are parasitic contributions from the overlap of the OECT electrodes.[1] Previous investigations suggested that the slower response time of OECTs with larger interdigitated electrodes was a result of the additional parasitic capacitance from the greater overlap between the gate electrode and the source and drain electrodes. Furthermore, the pinch-off voltages of these devices suggested that the dependence was primarily on channel thickness.[22] Therefore, the lower pinch-off voltages in the large comb
devices suggested that the OECT active layers were thinner than those of the small spiral. The differing thicknesses were thought to originate from the variances introduced in the photolithographic fabrication of the OECTs. These results imply that even with a thinner active layer in the large comb geometry, the bandwidth was significantly restricted by the parasitic contributions of the electrode overlap. This advises that the design of OECT amplifiers would benefit from minimizing contributions from parasitic capacitance.

The amplifier’s simulated temporal response was dependent only on the OECT response time constant, thus the cut-off frequency identified in the simulated Bode characterization was dependent only on this parameter. The mean of the cut-off frequency modeled by the simulation was found to be within one standard deviation of the physical values measured across the five devices used for each geometry. This supported that the exponential response of the OECT with response time constant $\tau$, as associated with a cut-off frequency of $f_c = 1/2\pi\tau$, was sufficient to model the amplifier bandwidth. The difference between cut-off frequencies of the two geometries was statistically significant and their values provide limits on the type of bioelectric signals that can be measured. With a cut-off frequency of $\approx 1$ kHz, the large comb geometry could measure and amplify bioelectric signals such as the local field potential (LFP), ECoG, EMG, and electrocardiography (ECG). The small spiral geometry, with a cut-off frequency of $\approx 5$ kHz, could amplify extracellular bioelectric activity of higher frequency.

The measurements of total harmonic distortion showed that the devices provide low distortion effects from inputs with amplitudes of $\approx 500$ $\mu$V. The difference in the lower limit of the THD between the OECT geometries was not statistically significant. The input-referred noise gave values of $\approx 100$ $\mu$V$_{\text{RMS}}$ in the bandwidth of the OECT, thus the noise became considerable at amplitudes lower than 500 $\mu$V. With no significant difference in the values of input-referred noise between the geometries and such noise dominating at low amplitudes, the lower limits of THD were expected to take similar values. Furthermore, the lower limit suggests that the OECT linear amplifier would be best implemented in unipolar recordings as opposed to bipolar recordings for electrophysiology. For the simulated upper limit, and thus for the full input dynamic range, the small spiral geometry gave a greater input range to a statistically significant level. However, the values for both geometries were greater than typical extracellular electrophysiological voltages, so both would still be appropriate for such measurements.

As shown in Table 2, the difference in RMS voltage noise between the two OECT geometries was not statistically significant when considering frequency bands between 0.5 Hz and 5 kHz. Relative to electrophysiological levels, the input-referred noise values imply that the measurable activity types are LFP, EMG, and ECG. This is assuming that there is no frequency filtering applied to reduce the level of noise. The distribution of noise with frequency followed a $1/f$ relationship, as shown in Figure 4c. The $1/f$ noise has previously been observed in OECT devices, which was suggested to originate from charge fluctuations in the bulk of the organic channel. It was previously observed that OECTs produce noise on the scales presented in Table 2 when not subject to frequency filtering. This suggests that the OECT device was the dominant source of noise in the amplifier topology. With no significant difference in noise between the OECT geometries, the results indicated that the noise was independent of the electrode overlap, supporting other investigations. However, it was also previously observed that OECT noise scaled inversely to channel area, which was found using varying parallel-plate electrode geometries. The channel areas varied by a factor of $\approx 6$, though the devices in this investigation utilized interdigitated electrode geometries. This suggests that further investigation is needed into the noise scaling relationships of interdigitated electrode architectures.

It was observed that the input-referred noise began to reduce relative to the $1/f$ behavior after 10 kHz, as illustrated in the subplot of Figure 4c. When considering both DC and AC behavior, the gate-ground voltage is dependent on the potential difference across the source resistor and bypass capacitor. For the values of $R_S$ and $C_I$ used in this investigation, the AC response of the amplifier was expected to only become evident for input frequencies approaching the 10 kHz-scale, which was greater than the bandwidth of the OECTs. For these frequencies, the impedance of the source resistor and bypass capacitor reduces, varying the potential difference across the source resistor and thus shifting the gate-ground voltage. The amplifier therefore operated around a different gate-ground voltage for these frequencies. As such, variations in gain, and thus input-referred noise, were expected. As evident in Figure 4c, the small spiral input-referred noise for frequencies at, and greater than, 10 kHz had values lower than those expected from extrapolating the lower-frequency $1/f$ noise. Therefore, the operating point of higher frequencies provided a lower gain, resulting in the attenuation of input voltages on the 10 kHz-scale. The bypass capacitor could thus be interpreted as a low-pass filter.

With a higher IP3 input power, an input of greater amplitude is required to equalize the third-order harmonics with the fundamental components, indicating a more linear device response. The physical measurements of the IP3 point did not give a statistically significant preference to either geometry. However, the simulation showed that the small spiral geometry had a greater linearity to a statistically significant level. The intercept point of the simulation indicated that harmonics became significant at input powers less than 10 dB, whereas extrapolation of the physical characterization suggested such input powers approaching 100 dB. A voltage resolution-limited simulation provided the expected gradient of 3 and agreed with physical measurements in showing that there was no clear linear increase in output power of third-order harmonics for input powers less than $-40$ dB (see Figure S8d, Supporting Information). However, physical measurements included input powers at $-25$ dB. At this input power, all simulations predicted a linear increase in third-order harmonics with a gradient of 3, which was not physically observed. The inability to identify a clear linear increase in all physical characterizations of the ten OECT amplifiers suggests that all circuit topologies provided an exceedingly high linearity within the range of voltages used in characterization. As the voltages were typical of the electrophysiological spectrum, this shows that the circuit topology was ideally suited to amplify such signals.
It is clear that the OECT amplifier presented in this work achieves low THD levels and high IP3 values, as expected from a good and practically useful amplifier. It should be stressed that a high IP3 value is a very desirable feature when designing high-performance and noise-robust biopotential acquisition systems. In hospital wards with multi-frequency noise sources and in applications where concurrent neural sensing and electrical stimulation is required, for example in deep brain stimulation and spinal cord stimulation setups, the nonlinearity of the amplifying system must be minimized to avoid artifact coupling with the electrophysiological measurements through intermodulation.[26]

2.6. Linear Amplification of Neural Signals

To further test the efficacy of the amplifier for electrophysiological signals, prerecorded bioelectric voltages were used as the gate input. The signals covered the entire extracellular electrophysiological spectrum and were previously recorded from patients (with full research ethics committee approvals) as part of other studies. Recordings were made using approved wired instruments and were taken from raw anonymized data. The signals were taken from measurements of ECoG, spreading depolarizations (SDs), and fasciculations from EMG. ECoG signals are recorded in patients with severe acute brain injuries to allow the detection of slow-moving mass SDs that are associated with poor patient outcomes.[31] EMG signals are used for patients with motor neuron disease to detect high-frequency fasciculations, which are predictive of disease progression.[32]

EMG fasciculations are associated with the discharge of motor units and thus have the action potential voltage–frequency characteristics of 10–100 µV-scale amplitudes with frequencies on the 0.1–1 kHz-scale.[13] The prerecorded ECoG data provided amplitudes of 0.1–1 mV with frequencies of 10–100 Hz. The pre-recorded SD data exhibited 10 mV-scale amplitudes and 10 kHz-scale frequencies.[14] The EMG fasciculation recordings contained frequency components greater than the upper bandwidth limit of the large comb OECTs. Therefore, the small spiral devices were used as the transistor in the circuit for this amplification due to the much greater bandwidth.

Amplified outputs of the 3 electrophysiological signals are shown in Figure 5. Frequency filters were applied to each signal depending on the frequencies typically dependent in that type. The gate input and amplified output were both subject to an identical bandpass filter, allowing signal comparison within the same frequency range. Notch filters were only applied to the amplified output and were present at 50 Hz and higher odd harmonics. The SD signals were bandpass filtered between 5 mHz and 500 Hz, the ECoG between 1 Hz and 500 Hz and the EMG fasciculations between 10 Hz and 5 kHz.

The amplifier response to the EMG fasciculations is presented in Figure 5a. Incremental variations approaching as low as 100 µV were resolvable and were amplified to 200 µV outputs. It was evident that noise started to dominate the response at those amplitudes. The values of RMS voltage noise given in Table 2 suggested noise levels were ≈100 µV. Furthermore, at such amplitudes there were significant contributions from higher harmonics, as highlighted in Figure 4a. The utilization of notch and bandpass filters allowed the resolution of features with amplitudes of 100 µV. However, features of smaller voltage amplitudes were not resolvable. In extracellular environments, the OECT’s mechanism for detecting electrophysiological signals has been shown to provide superior signal-to-noise ratios compared to electrodes within certain frequency bands.[3] Therefore, the effective gate voltage is expected to be greater than the potential differences recorded using electrodes. As such, if utilizing the OECT linear amplifier where extracellular currents are measured using electrochemical dedoping (as given in Figure S2, Supporting Information), then the constraints on the lower limits of voltage resolution may be less strict. Nevertheless, reductions in noise would increase efficacy.

The response from the ECoG input, as shown in Figure 5b, was successful in resolving and amplifying the features present in the signal. This was expected because the potential differences were within the frequency bandwidth, greater than the RMS voltage noise and associated with THD values approaching 1%. The OECT amplifier was also able to accurately reproduce and amplify the features of the SD input, as illustrated in Figure 5c. The low-frequency spreading depolarization is identifiable in the amplifier output. With such large amplitudes of such events, the relative level of noise is negligible. Of the amplifier characteristics, the input dynamic range takes greater consideration for these bioelectric signals. With potential differences taking a range within 20 mV and the simulated input dynamic range having a width of approaching 100 mV, the circuit was able to amplify the input signal with high linearity.

**Figure 5.** Response of a 35 nm small spiral OECT common-source linear amplifier with $R_1 = 1 \, \text{k} \Omega$, $R_2 = 55 \, \Omega$, $R_3 = 180 \, \Omega$, $R_D = 600 \, \Omega$, $V_{DD} = -2 \, \text{V}$, and $C_1 = 100 \, \text{nF}$ and with OECT characteristics of $G = 6.06 \, \text{mS}$, $V_F = 0.53 \, \text{V}$, and $\tau = 32 \, \mu\text{s}$ to a prerecorded gate voltage input of a) EMG fasciculations (EMG fasc.), b) ECoG data, c) a spreading depolarization (SD). The ideal output was given as the product of the input signal with the amplifier gain.
3. Conclusion

The linear amplifier circuit was successfully implemented with an OECT as the amplifying element. With the physical limitations of the OECT under consideration, viable combinations of voltage and resistor values were found by using the output characteristics in a DC load-line analysis. This gave sufficient information to determine amplifier parameters such as gain, which allowed further investigation to correlate potential amplifier performance with OECT parameter values. Using the drain-source voltage as the amplifier output, it was found that lower pinch-off voltages, which were associated with thinner OECT active layers, gave higher gain values.

To simulate the dynamics of the circuit, an equation was identified which showed a very good fit to OECT output characteristics across the linear, nonlinear, and saturation regions. This allowed the determination of an amplifier transfer curve, showing the output voltage given a gate-ground input voltage. The equation was used to simulate the response of the common source linear amplifier given an arbitrary, low frequency gate-ground voltage input.

In determining that the 35 nm devices could give greater gain values than the 70 nm devices, the 35 nm small spiral and large comb OECTs were physically implemented into a common-source linear amplifier. The voltage and resistor values were chosen based on those previously found in the DC load-line analysis. Previous literature on voltage amplification with OECTs had not included the characterization of a number of figures of merit, and thus comprehensive testing was carried out for such an analysis. In particular, the total harmonic distortion, intermodulation distortion, input-referred noise, and bandwidth were accurately determined. Furthermore, the circuit topologies were simulated with the same inputs used to calculate the figures of merit, allowing for comparison to the physical measurements and insight into amplifier behavior outside the electrophysiological bounds.

The figures of merit were compared between the device types based on the mean and standard deviation of 5 devices tested per type. Using a t-test of independence assuming unequal variances, it was found that the cut-off frequency and gain were significantly different. Physical measurements found no significant difference in the lower bound for 1% THD, though simulations suggested that input dynamic range for the small spiral geometry was greater than the large comb geometry to a significant degree. There was no significant difference in the level of noise present in the two geometries and the corner frequency could not be identified in the 50 kHz frequency range. The physical measurements of intermodulation distortion did not show preference between OECT types, though the simulations suggested that the small spiral geometry had superior linearity.

Finally, prerecorded electrophysiological signals were inputted into the gate of the small spiral device. With the application of frequency filters, the primary features of EMG fasciculations could be resolved and amplified. The efficacy to amplify bioelectric signals was greater for the ECoG and SD inputs, which was expected when comparing the RMS voltage noise to the magnitude of voltage changes expected in the different signal types.

In this investigation, the OECT common source linear amplifier was comprehensively characterized to determine the figures of merit and has shown capability to amplify electrophysiological signals. The biocompatibility of the OECT, derived from its use of organic and biologically inert materials, provides potential implementation into amplification local to physiological environments. With the circuit topology allowing for a more fine-tuned amplification, this work has illustrated the capabilities of bioelectronic voltage amplifiers.

4. Experimental Section

OECT Fabrication and Characterization: The devices used in this investigation were fabricated identically to those used previously.[22] PET substrates with a thickness of 125 µm were cleaned by alternating sonication in acetone and IPA and were then subject to an oxygen plasma. Thermal evaporation was used to layer 15 nm Cr and either 200 or 300 nm Au onto the PET substrates. Photolithography and etching with Cr and Au etchants were used to pattern the source and drain electrodes. A Karl Suss MJB-3 mask aligner, S1805 photoresist, and MF-26A developer were used. The encapsulation layer consisted of a 1 µm thick layer of SU-8, which was spin-coated and subsequently patterned using photolithography. The development process gave the final encapsulation geometry, and the developer propylene glycol monomethyl ether acetate from Sigma-Aldrich was used. The substrate was baked for 20 min at 120 °C. The PEDOT:PSS composition consisted of 91.7% Heraeus Clevisoph PH1000 PEDOT:PSS, 8% ethylene glycol, and 0.3% dodecyl-benzene-sulfonic-acid. The PEDOT:PSS composition was spin-coated to a thickness of either 35 or 70 nm and patterned using photolithography and etching with an oxygen plasma ash. After baking for 75 min at 150 °C, the substrates were submersed in DI water for 15 min. Finally, Ringer’s solution was used for the electrolyte and the gate electrode was an Ag/AgCl pellet electrode.

Devices were characterized by connecting the electrodes through a probe station into a Keysight B2902a precision source/measure unit (SMU). The source and drain electrodes were connected to one SMU and the source and gate to the other. Characterization was performed using the quick IV software and was analyzed using custom python code. The output characteristics were found by sweeping the drain-source voltage from ~0.6 to ~0.6 V for a set of gate-source voltages from ~0.2 to ~0.6 V in 0.1 V steps. The current at the drain electrode, the drain-source voltage and the gate-source voltage were measured during the characterization.

Linear Amplifier Characterization: To perform the characterization of the linear amplifier, data acquisition was performed using the Labchart software with PowerLab 16/3S. The voltages were sampled at 100 kHz across all channels used. An AC coupling filter was implemented within the LabChart software on all characterizations. For the measurement of THD and frequency response, the Agilent 33220a was used as the waveform generator. For the IMD measurements, two CW INSTEK AFG-2125 waveform generators were used. Voltage attenuators were incorporated when needed at the lower gate voltage values. The inputted THD waveforms had amplitudes which were iterated but had a constant frequency of 5 Hz. Similarly, the IMD waveform was a superposition of two sinusoids of equal amplitude (amplitudes were iterated) at the frequencies of 4.9 and 5.1 Hz. For each waveform inputted in the THD and IMD characterizations, the output voltage was recorded for 1 min. To measure the input-referred noise, the input voltage was grounded, and the output voltage was recorded for 1 min. For the characterization of frequency response, two frequency sweeps were performed: from 1 to 100 Hz and from 100 Hz to 10 kHz. The period of each frequency sweep was 100 s and the amplitude of the sinusoid inputted was 10 mV. The gain was determined by the results of the THD characterization.

Bioelectric Testing: The gate electrode of the OECT in the linear amplifier configuration was connected to a Keysight 33220A waveform generator.
The waveform generator was preloaded with electrophysiological signals as previously described.\[^{[29]}\] Voltage attenuators were used to reduce the magnitude of the output voltage below the minimum of 20 mV. The voltage response was sampled at 100 kHz using Labchart software with PowerLab 16/35.

**Supporting Information**

Supporting Information is available from the Wiley Online Library or from the author.

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**Conflict of Interest**

The authors declare no conflict of interest.

**Data Availability Statement**

Research data are not shared.

**Keywords**

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