Energy-Harvesting Piezoelectric-Powered CMOS Series Switched-Inductor Bridge

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Abstract—Tiny piezoelectric transducers transform a very small fraction of ambient energy into electrical power. Power losses and breakdown voltages unfortunately limit how much of this power reaches the microsystem. The piezoelectric harvester presented here operates at the maximum power point (MPP) without the separate MPP stage that typical piezoelectric harvesters need. As a result, the system loses less power and occupies less space. Plus, with only four transistors, the charger increases the voltage with which the piezoelectric current generates power. This way, the harvester draws more power from vibrations. By pre-charging the transducer between half cycles, the 0.81–1.2-µm CMOS system outputs 0.7–49 µW of the 1.2–55 µW that a vibrating 16.8-nF transducer avail at 120 Hz. This is up to 6.8 \times more power than a lossless diode bridge can harness and as much as the best recycling bridge can output, but with four fewer power switches and without a separate maximum power-point stage.

Index Terms—Energy harvester, piezoelectric charger, switched-inductor bridge, maximum power point (MPP), damping force, synchronized discharges, power index, synchronized switch damping/harvesting on inductor (SSDI and SSHI).

NOMENCLATURE

\( i_{PZ} \) Piezoelectric current [A]
\( i_{PZ}(PK) \) Peak piezoelectric current [A]
\( C_{PZ} \) Piezoelectric capacitance [F]
\( C_{PZ(XFR)} \) Piezoelectric capacitance during energy transfers between half cycles [F]
\( E_{PC} \) \( C_{PZ} \)'s energy at \( v_{PC} \) [J]
\( E_{PK} \) \( C_{PZ} \)'s energy at \( v_{PZ(PK)} \) [J]
\( E_{XFR} \) Energy \( C_{PZ} \) transfers between half cycles [J]
\( i_L \) Inductor current [A]
\( P_{LOSS} \) Power losses [W]
\( P_O \) Output power [W]
\( P_{PZ} \) Drawn piezoelectric power from transducer [W]
\( P_{PZ(BRDG)} \) Power generated by a lossless diode bridge [W]
\( P_{PZ(MAX)} \) The maximum power that \( i_{PZ} \) can supply [W]
\( V_{BD} \) CMOS breakdown voltage [V]
\( V_{PC} \) Pre-charge voltage [V]
\( V_{PZ} \) Piezoelectric voltage [V]
\( V_{PZ(PK)} \) Peak piezoelectric voltage [V]
\( \Delta V_{PZ(OC)} \) Transducer's open-circuit voltage [V]

\( \eta_{i(PZ)} \) Drawn power index [W/W]
\( \eta_{i(O)} \) Output power index [W/W]

I. PIEZOELECTRIC ENERGY-HARVESTING MICROSYSTEMS

Sensors can monitor, process, store, and communicate data about their surrounding environment that save money, energy, and lives. In practice, many existing and emerging applications that can benefit from this technology cannot fit connectors or ports. Sensors should therefore be small and self-sustaining [1]–[2]. With so much functionality, however, tiny onboard batteries can drain quickly [3]. Luckily, vibrations are often available and abundant [4]–[5], so ambient energy in motion can replenish the energy that these small batteries supply.

Small transducers only draw 1% to 5% of the energy available in motion [6]. Piezoelectric generators are more popular than electromagnetic and electrostatic transducers are because they output 1.3 \times to 3 \times more power under similar constraints and conditions [7]. But since power-supply circuits burn power, the system receives less power than this [8]. In some cases, still less power is available because transistor breakdown voltages limit the voltage with which harvesters can draw power [9]. This is why energy harvesters should operate at their maximum power point (MPP) [10].

Energy-harvesting systems powered with piezoelectric transducers normally include MPP features for this purpose. In Fig. 1, for example, the charger draws power from a tiny piezoelectric generator. Since small transducers output little power, a battery \( v_B \) is used to cache power. When \( v_B \) holds enough energy, a power supply draws power from \( v_B \) to feed the sensors, interface amplifiers (\( A_V \)), analog–digital converters (ADC), digital-signal processors (DSP), memory, and power amplifiers (PA) in the system. The purpose of the maximum power-point tracker (MPPT) is to adjust the charger so it outputs the highest power possible.

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Fig. 1. Piezoelectric-powered energy-harvesting microsystem.

The principal aim of the charger is to draw the highest power possible with the least losses and under the least constraints. The four-transistor series switched-inductor bridge
presented here does this by pre-charging the piezoelectric transducer to a level that is adjustable. This way, without another power stage, the charger can operate at the MPP with few power-consuming transistors switching across a vibration cycle. To understand the details and benefits of this bridge, Sections II–V describe operation, discuss output power, and assess and compare performance with the state of the art.

II. PROPOSED SERIES SWITCHED-INDUCTOR BRIDGE

A. Piezoelectric Transducer

Charge centers in a piezoelectric cantilever misalign when deformed [11]. The voltage potential that this establishes across the capacitance \( C_{PZ} \) of the transducer induces charge flow. So when a transducer vibrates without an electronic load, the resulting current \( i_{PZ} \) charges and discharges \( C_{PZ} \) across \( \Delta v_{PZ(OC)} \). The transducer therefore delivers and recovers the energy that \( C_{PZ} \) receives.

But when a harvester extracts the energy that \( C_{PZ} \) receives, less energy is available to push the piezoelectric cantilever. As a result, displacement distance decreases, and in consequence, so does \( i_{PZ} \). Electrical energy in \( C_{PZ} \) is so low with respect to mechanical energy in the system, however, that draining \( C_{PZ} \)'s energy has nearly no effect on the moving cantilever or the charge that \( i_{PZ} \) supplies [6], [12]–[21]. In other words, \( i_{PZ} \) is nearly independent of the damping force with which the harvester draws power. This means that drawing more power nearly independent of the damping force with which the harvester draws power. This means that drawing more power from the transducer has little effect on \( i_{PZ} \).

B. Operation

The first basic objective of the charger proposed is to draw the power that \( C_{PZ} \) collects across each half cycle. For this, the switched-inductor bridge in Fig. 2 drains \( C_{PZ} \) between half cycles. This is why \( C_{PZ} \)'s voltage \( v_{PZ} \) in Fig. 3 falls to zero at 4.2, 8.4, and 13 ms. The second aim of the charger proposed is to raise the voltage \( v_{PZ} \) with which \( C_{PZ} \) collects charge. This way, with higher \( v_{PZ} \), \( i_{PZ} \) generates more power \( P_{PZ} \) or \( i_{PZ} v_{PZ} \). The bridge does this by recycling some of the energy drawn to pre-charge \( C_{PZ} \) to \( v_{PC} \) before every half cycle begins (at 4.2, 8.4, and 13 ms). Therefore, \( v_{PZ} \) starts at \( v_{PC} \) and rises to \( v_{PC} + \Delta v_{PZ(OC)} \) across each half cycle. With a higher \( v_{PZ} \) (that is always greater than \( v_{PC} \)), the system draws more power from motion than without \( v_{PC} \), which means the damping force against vibrations is greater.

![Diagram of the proposed piezoelectric-powered series switched-inductor bridge.](image)

**Fig. 2.** Proposed piezoelectric-powered series switched-inductor bridge.

To start this, all but the bottom ground switch \( M_{GB} \) open across \( i_{PZ} \)'s positive half cycle, so the bridge open-circuits and \( i_{PZ} \) charges \( C_{PZ} \) to \( v_{PZ(PK)} \) across 0.1–4.2 ms in Fig. 3. The top ground switch \( M_{GT} \) then closes to drain \( C_{PZ} \) into the transfer inductor \( L_{X} \). As \( L_{X} \) energizes, \( v_{PZ} \) falls to zero and \( L_{X} \)'s current \( i_{L} \) in Fig. 4 increases (across \( t_{E} \) at 4.167–4.170 ms). \( M_{GB} \) and \( M_{GT} \) remain closed for another short interval \( t_{D1} \) to partially drain \( L_{X} \) into \( C_{PZ} \), and that way, pre-charge \( C_{PZ} \) in the negative direction. After that, \( M_{GT} \) opens and the top battery switch \( M_{BT} \) closes across \( t_{D2} \) to deplete \( L_{X} \) into the battery \( v_{B} \) and \( C_{PZ} \). So \( v_{B} \) charges after the positive half cycle, like Fig. 3 shows, and \( C_{PZ} \) charges to \(-v_{PC}\) before the negative half cycle.

![Graph showing measured piezoelectric and battery voltages across cycles.](image)

**Fig. 3.** Measured piezoelectric and battery voltages across cycles.

![Graph showing measured inductor current at the end of a positive half cycle.](image)

**Fig. 4.** Measured inductor current at the end of a positive half cycle.

![Graph showing measured inductor current at the end of a negative half cycle.](image)

**Fig. 5.** Measured inductor current at the end of a negative half cycle.

The process repeats across negative half cycles. In this case, all but \( M_{GT} \) open across 4.2–8.4 ms in Fig. 3, so \( i_{PZ} \) charges \( C_{PZ} \) from \(-v_{PC}\) to \(-v_{PZ(PK)} \), \( M_{GB} \) then closes at 8.320 ms in Fig. 5 to drain \( C_{PZ} \) into \( L_{X} \) (across \( t_{D1} \)) and a little longer (across \( t_{D2} \)) to partially drain \( L_{X} \) into \( C_{PZ} \). After that, \( M_{GB} \) opens and \( M_{BB} \) closes (across \( t_{D2} \)) to deplete \( L_{X} \) into \( v_{B} \) and \( C_{PZ} \). This way, \( v_{B} \)
receives charge and $C_{PZ}$ pre-charges to $v_{PC}$ in Fig. 3.

The entire sequence, which Table I summarizes, repeats every cycle to exhibit the steady-state behavior shown in Fig. 3. Figure 6, for example, shows how the system charges 270 nF across 2.7–4.2 V across 40 ms in approximately 50-mV increments. This voltage range corresponds to that of thin-film Li-Ion batteries [22]. Total charge time ultimately depends on the capacity of the battery and vibration strength.

<table>
<thead>
<tr>
<th>$i_{PZ}$’s State</th>
<th>$M_{GT}$</th>
<th>$M_{GA}$</th>
<th>$M_{BT}$</th>
<th>$M_{BA}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Across + ½ Cycle</td>
<td>Open</td>
<td>Closed</td>
<td>Open</td>
<td>Open</td>
</tr>
<tr>
<td>+/– Transition $t_1$</td>
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<td>Open</td>
<td>Open</td>
<td>Open</td>
</tr>
<tr>
<td>+/– Transition $t_2$</td>
<td>Open</td>
<td>Closed</td>
<td>Open</td>
<td>Open</td>
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<tr>
<td>Across – ½ Cycle</td>
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<td>Closed</td>
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<tr>
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<td>Open</td>
</tr>
<tr>
<td>–/+ Transition $t_4$</td>
<td>Open</td>
<td>Open</td>
<td>Closed</td>
<td>Closed</td>
</tr>
</tbody>
</table>

*+* and *–* refer to $i_{PZ}$’s polarity.

**TABLE I: SWITCHING SEQUENCE**

Fig. 6. Measured charging profile.

**C. Power Transistors**

The switches require gate-drive power $P_G$ to switch and ohmic power $P_R$ to conduct. Since gate capacitance $C_G$ and channel resistance $R_{CH}$ increase with channel length $L_{CH}$, $L_{CH}$ should be minimum length $L_{MIN}$. In this case, the $L_{MIN}$ that can withstand 5.5 V is 1.2 μm for NFETs and 0.81 μm for PFETs. But since $C_G$ increases and $R_{CH}$ decreases with wider channels, channel widths $W_{CH}$ in Fig. 3 balance $P_C$ and $P_R$ [23] when the system draws 30 µW from the transducer, which here corresponds to the most probable vibration strength. Optimizing $W_{CH}$ for the most likely condition saves energy, and in consequence, outputs the most power.

**D. Drivers**

Three inverters with transistors of increasing dimensions drive each switch. The first inverter in the driver is minimum size, the second is 5× larger than the first, and the third is 5× larger than the second. Although 2.67× is optimal for shortest propagation delay [24], a higher gain reduces the number of inverter stages, and as a result, the shoot-through power that they consume [25].

Since NFETs connect to ground and ground is the lowest potential in the circuit, NFETs can open and close with the voltage that a lithium-ion battery can supply: 2.7–4.2 V [22]. This is why the supplies for NFET drivers are $v_B$ and ground. For PFETs to open, their gates must charge to the highest terminal voltage. PFETs here connect to $v_B$ and switching nodes $v_{SWT}$ and $v_{SWB}$, which rise to a level $v_{PZ(PK)}$ (in Fig. 3) that depends on pre-charge voltage $v_{PC}$ and vibration strength $\Delta v_{PZ(OCC)}$. This means, $v_{SWT}$ and $v_{SWB}$ may or may not surpass $v_B$, so $v_B$ is not a good supply for PFET drivers.

**E. Maximum-Supply Selector**

The purpose of the maximum-supply selector block Max in Fig. 2 is to establish the highest supply $v_{MAX}$ with which PFET drivers can open P-type switches. For this, cross-coupled PFET pair $M_{1A}$–$M_{1B}$ in Fig. 7 selects and connects the higher of $v_{SWT}$ and $v_{SWB}$ to $v_{O1}$: $M_{1A}$ connects $v_{SWT}$ to $v_{O1}$ when $v_{SWB}$ is below $v_{SWT}$ by more than a PFET threshold voltage $|v_{TP}|$ and $M_{1B}$ connects $v_{SWB}$ to $v_{O1}$ when the opposite is true. $M_{2A}$–$M_{2B}$ similarly selects and connects the higher of the resulting $v_{O1}$ and $v_B$ to $v_{MAX}$. So together, $v_{MAX}$ is the highest of the three voltages:

$$v_{MAX} = \text{Max} \left\{ v_{O1} \right\} = \text{Max} \left\{ v_{SWT} \right\} = \text{Max} \left\{ v_{SWB} \right\}.$$  (1)

Fig. 7. Prototyped maximum-supply-selector.

$v_{MAX}$ in Fig. 8, for example, connects to $v_B$ at 1.8–3.4 ms because $v_B$’s 2.9 V exceeds $v_{SWT}$’s 0–2.9 V and $v_{SWB}$’s 0 V. $v_{MAX}$ connects to $v_{SWT}$ at 3.4–5.8 ms because $v_{SWT}$’s 2.9–4.5 V similarly surpasses $v_B$’s 2.9 V and $v_{SWB}$’s 0 V. $v_{MAX}$ then connects to $v_{SWB}$ at 7.4–10 ms because $v_{SWB}$’s 2.9–4.5 V is greater than $v_B$’s 2.9 V and $v_{SWT}$’s 0 V.

**Fig. 8. Measured maximum-supply-selector waveforms in steady state.**

Although $M_{1A}$, $M_{1B}$, $M_{2A}$, and $M_{2B}$ do not conduct as much current as power NFETs and PFETs in Fig. 2, they still supply...
the charge that PFET gates in Fig. 2 need to switch between states. This is why their channel lengths (in Fig. 7) are the shortest possible that can withstand 5.5 V. Their channel widths ensure M1_A, M1_B, M2_A, and M2_B drop less than 100 mV when charging PFET gates in Fig. 2.

Cross-coupled PFETs do not close when their terminal (gate) voltages are within a threshold \(|v_{TP}|\) of one another. Here, however, M1_A–M1_B's \(v_{SWT}\) and \(v_{SWB}\) are within a \(|v_{TP}|\) only between half cycles across 2–4-\(\mu\)s transitions, when M_GT and M_GB connect \(v_{SWT}\) and \(v_{SWB}\) to ground. This is not a problem because \(v_B\) is much greater than \(v_{SWT}\) and \(v_{SWB}\) during this time, so M2_B connects \(v_B\) to \(v_{MAX}\).

M2_A–M2_B's \(v_{OH}\) and \(v_B\) are within a \(|v_{TP}|\) only when \(C_{PZ}\) charges high enough to be within a \(|v_{TP}|\) of \(v_B\), which does not always happen. If \(v_{PZ(PK)}\) is well above \(v_B\), for example, \(v_{SWT}\) and \(v_{SWB}\) are within a \(|v_{TP}|\) of \(v_B\) only when \(v_{SWT}\) and \(v_{SWB}\) cross \(v_B\). This is not a problem because transitions are short and PFET gates do not need charge halfway across transitions. But even if \(v_{PZ}\) peaks within a \(|v_{TP}|\) of \(v_B\), M1_A–M1_B and M2_A–M2_B's combined bulk-to-substrate capacitance \(C_{BULK}\) is large enough to hold and supply the charge that PFET gates in Fig. 2 need across this short interval. Although adding capacitance helps, \(C_{BULK}\) adds to \(C_{PZ}\) when \(v_{SWT}\) or \(v_{SWB}\) is higher than \(v_B\), so \(C_{BULK}\) steers \(i_{PZ}\) away from \(C_{PZ}\). In other words, adding capacitance sacrifices energy that would otherwise reach \(C_{PZ}\).

The worst-case condition occurs when \(v_{SWT}\) and \(v_{SWB}\) cross \(v_B\) within half cycles (at, for example, 3.4, 7.4, and 11.4 ms in Fig. 8). Across these millisecond crossings, \(v_{MAX}\) is a \(|v_{TP}|\) below \(v_{SWT}\) and \(v_{SWB}\). With \(|v_{TP}|\) of gate drive, however, battery PFETs M_BT and M_BB in Fig. 2 operate in weak resistive in weak inversion. So although their effect is to leak \(v_B\) to ground and to leak \(C_{PZ}\) to \(v_B\), leakage is low because these PFETs are very resistive in weak inversion.

### III. OUTPUT POWER

#### A. Drawn Piezoelectric Power

The system draws between half cycles the energy that \(C_{PZ}\) collects across half cycles. The energy that \(C_{PZ}\) needs (\(E_{PC}\)) to pre-charge to \(v_{PC}\) essentially cycles between \(C_{PZ}\) and \(L_X\). So of the energy drawn when \(v_{PZ}\) peaks (\(E_{PK}\)), the transducer supplies the difference \(E_{PK} - E_{PC}\) every half cycle and twice that difference \(2(E_{PK} - E_{PC})\) every full cycle \(f_{VIB}\).

Since \(i_{PZ}\) charges \(C_{PZ}\) across \(\Delta V_{PZ(OC)}\) every half cycle, \(v_{PZ}\) increases from \(v_{PC}\) (at 4.2, 8.4, and 13 ms in Fig. 3) to \(v_{PC} + \Delta V_{PZ(OC)}\), \(C_{PZ}\)'s peak voltage \(v_{PZ(PK)}\), \(C_{PZ}\)'s peak energy \(E_{PK}\), or \(0.5C_{PZ}v_{PZ(PK)}^2\) is therefore \(0.5C_{PZ}(v_{PC} + \Delta V_{PZ(OC)})^2\). But after subtracting \(E_{PC}\)'s \(0.5C_{PZ}v_{PC}^2\), drawn piezoelectric power \(P_{PZ}\) reduces to

\[
P_{PZ} = \frac{E_{PK} - E_{PC}}{f_{VIB}}
\]

\[
= 2 \left(0.5C_{PZ}v_{PZ(PK)}^2 - 0.5C_{PZ}v_{PC}^2\right) f_{VIB}
\]

\[
= 2 \left(0.5C_{PZ} \left(v_{PC} + \Delta V_{PZ(OC)}\right)^2 - v_{PC}^2\right) f_{VIB}
\]

\[
= C_{PZ} \left(\Delta V_{PZ(OC)}^2 + 2v_{PC}\Delta V_{PZ(OC)}\right) f_{VIB}
\]

where \(f_{VIB}\) is the frequency of vibrations.

The underlying assumption here is that \(P_{PZ}\)'s (loading) effect on motion is negligible, which is the case for small piezoelectric devices [6], [12]–[21]. As a result, \(i_{PZ}\) is nearly unaffected by how much power the harvester draws with \(P_{PZ}\), which is another way of saying the transducer can source more power than it actually supplies. This is why pre-charging \(C_{PZ}\) is so important, because \(v_{PC}\) raises the voltage \(v_{PZ}\) with which \(i_{PZ}\) generates \(P_{PZ}\). And with a higher \(v_{PZ}\), the transducer sources more power.

#### B. Transducer Losses

The mechanical properties of the transducer dictate how much piezoelectric capacitance \(C_{PZ}\) appears across its terminals [26]. In the case of the unit tested, \(C_{PZ}\) in Fig. 9 is 16.8 nF at the vibration frequency \(f_{VIB}\), which here is 120 Hz. This is the capacitance that collects piezoelectric charge across half cycles.

![Fig. 9. Measured piezoelectric capacitance across frequency.](image-url)

Interestingly, the effective capacitance falls as frequency climbs. This is not unreasonable because uneven distribution of so much series resistance \(R_{PZ}\) steers more current into capacitive components with lower resistance. This is why 14.5 nF of the 16.8 nF available activate when transferring energy at 70 kHz.

Here, 40–80 kHz corresponds to the 2–4 \(\mu\)s that \(L_X\) needs to transfer \(C_{PZ}\)'s energy between half cycles. In other words, a lower average capacitance \(C_{PZ(XFR)}\) of 14.5 nF transfers between half cycles the energy that the higher counterpart \(C_{PZ}\) collects across half cycles. \(C_{PZ(XFR)}\) therefore transfers the pre-charging charge \(q_{PC}\) required to pre-charge \(C_{PZ}\) to \(v_{PC}\):

\[
q_{PC} = C_{PZ}v_{PC} = C_{PZ(XFR)}v_{PZ}.
\]

Except, this same \(q_{PC}\) establishes a higher voltage \(v_{PC}'\) across \(C_{PZ(XFR)}\)'s lower capacitance. This is why \(C_{PZ}\)'s pre-charging level in Fig. 3 falls slightly at the beginning of every
half cycle, because $C_{PZ(XFR)}$’s $v_{PC}$ drops to $C_{PZ}$’s $v_{PC}$.

This is unfortunate because a linear rise in voltage produces a quadratic rise in energy that outpaces a linear fall in capacitance. In other words, $C_{PZ(XFR)}$ needs more energy ($E_{XFR}$) than $C_{PZ}$ requires ($E_{PC}$) to charge $C_{PZ}$ to $v_{PC}$. Series resistances in $C_{PZ}$ burn this difference $E_{XFR} - E_{PC}$ every half cycle and twice every full cycle $t_{VIB}$, so this loss $P_{CPZ}$ is

$$P_{CPZ} = 2 \left( \frac{E_{XFR} - E_{PC}}{t_{VIB}} \right)$$

$$= \left( C_{PZ(XFR)} v_{PC}^2 - C_{PZ} v_{PC}^2 \right) \frac{v_{PC}}{t_{VIB}}$$

$$= \left[ C_{PZ(XFR)} \left( \frac{C_{PZ} v_{PC}}{C_{PZ(XFR)}} \right)^2 - C_{PZ} v_{PC}^2 \right] \frac{v_{PC}}{t_{VIB}}$$

(4)

$$= C_{PZ} v_{PC}^2 \left( \frac{C_{PZ}}{C_{PZ(XFR)}} - 1 \right) \frac{v_{PC}}{t_{VIB}}$$

Note that $P_{CPZ}$ vanishes when $C_{PZ(XFR)}$ and $C_{PZ}$ match, when all capacitive components in the transducer transfer energy between half cycles. Also notice that $P_{CPZ}$ scales with $v_{PC}$, so $P_{CPZ}$ increases when $v_{PC}$ rises. Plus, the series resistance $R_{PZ}$ that conducts the power at 40–80 kHz burns ohmic losses $P_{RPZ}$. So imperfections in the transducer ultimately cost the system ohmic and dynamic losses $P_{RPZ}$ and $P_{CPZ}$.

C. Maximum Output Power

Loss Limit: Power losses $P_{LOSS}$ limit how much of the drawn piezoelectric power $P_{PZ}$ the system can deliver. Unfortunately, all components lose power. The transducer loses ohmic power $P_{RPZ}$ to $R_{PZ}$ and dynamic power $P_{CPZ}$ when pre-charging $C_{PZ}$. The inductor $L_X$’s series resistance $R_L$ also burns ohmic power $P_{RL}$. MOS transistors consume ohmic power $P_{MR}$ to conduct and need gate-drive power $P_{MG}$ to switch. Plus, drivers burn shoot-through power $P_{ST}$ when they transition. And although not nearly as much, power PFETs leak piezoelectric and battery power $P_{LK}$ when the switching nodes $v_{SWT}$ and $v_{SWB}$ cross $v_B$ and $C_{BULK}$ leaks piezoelectric power $P_{BL}$ away from $C_{PZ}$ when $v_{SWT}$ and $v_{SWB}$ surpass $v_B$. So of $P_{PZ}$, the battery $v_B$ receives $P_{PZ} - P_{LOSS}$, where $P_{LOSS}$ is

$$P_{LOSS} = P_{RPZ} + P_{CPZ} + P_{RL} + P_{MR} + P_{MG} + P_{ST} + P_{Lk} + P_{BL}. \quad (5)$$

Power-conversion efficiency $\eta_C$ is therefore the fraction of $P_{PZ}$ that all these losses $P_{LOSS}$ in the system avail with output power $P_O$:

$$\eta_C = \frac{P_{PZ} - P_{LOSS}}{P_{PZ}} = 1 - \frac{P_{LOSS}}{P_{PZ}}.$$

Notice that fractional losses $P_{LOSS}/P_{PZ}$ set this efficiency.

Since ohmic losses $P_R$ climb with $L_X$’s root–mean–squared conduction current $i_L(RMS)$ and $i_L(RMS)$ climbs with piezoelectric power $P_{PZ}$, losses $P_{RPZ}$, $P_{RL}$, and $P_{MR}$ increase with $P_{PZ}$. But while $P_{PZ}$ climbs linearly with $i_L(RMS)$, $P_R$ grows quadratically (with $i_L(RMS)^2R_{EQ}$) for resonant transfers ($t_E$ and $t_{D1}$ in Fig. 4 and Fig. 5) and cubically (with $i_L(RMS)^3R_{EQ}$) for battery transfers ($t_{D1}$ in Fig. 4 and Fig. 5) [20]. This means that, when vibration strength is low, an increase in $P_{PZ}$ exceeds the rise in $P_R$ to produce a net gain in $P_O$. Eventually, however, $P_R$’s quadratic-to-cubic loss outpaces $P_{PZ}$’s linear gain to the extent that $P_O$ falls. Plus, $P_{CPZ}$ also increases quadratically with pre-charging voltage $v_{PC}$. So even though $P_{PZ}$ rises monotonically across $v_{PC}$’s entire 0–4-V range in Fig. 10, $P_O$ maxes at 12 µW when vibration strength peaks $i_{PZ}$ to 12 µA and charges $C_{PZ}$ across 2.0 V.

As vibrations gain strength, $P_R$’s rise cancels $P_{PZ}$’s gain at higher power levels. This is why the maximum power point $P_{OMPP}$ in Fig. 11 increases with $i_{PZ(PK)}$ up to 21 µA. As this happens, $P_{RPZ}$, $P_{RL}$, and $P_{MR}$ rise. $P_{MG}$ and $P_{ST}$ do not increase with $i_{PZ(PK)}$ because gate-drive losses and shoot-through power do not scale with vibration strength. $P_{RPZ}$ and $P_{RL}$ dominate because $R_{PZ}$ and $R_L$ in small devices are much higher at 1–10 Ω [27]–[28] than MOS resistances, which engineers normally keep at milliohms [29]. In other words, scaling down the size of the transducer and the inductor produces the losses that dominate and limit $P_O$.

Since the power $P_{CPZ}$ that $C_{PZ}$ loses when pre-charging scales with $v_{PC}$ and $v_{PZ}$ increases with vibration strength in this region, $P_{CPZ}$ also increases with $\Delta v_{PZ(OC)}$ up to 3.5 V. In all, the system delivers 45%–91% of the 1.2–55 µW that the transducer supplies with $i_{PZ}$ and $v_{PZ}$ in $P_{PZ}$. Power-conversion efficiency $\eta_C$ is lower when vibrations are weak because switching losses $P_{MG}$ and $P_{ST}$, which do not scale with $\Delta v_{PZ(OC)}$, dominate when $P_{PZ}$ is low.

Breakdown Limit: Recall that $C_{PZ}$’s voltage $v_{PZ}$ is the voltage with which the piezoelectric current $i_{PZ}$ supplies power $P_{PZ}$. So for the same $i_{PZ}$, the transducer outputs more power when $v_{PZ}$

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**Fig. 10.** Measured power across $C_{PZ}$’s pre-charge voltage.

**Fig. 11.** Measured power and losses.

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**Table 1:**

<table>
<thead>
<tr>
<th>Pre-Charge Voltage $v_{PC}$ [V]</th>
<th>Power Output $P_O$ [W]</th>
<th>Losses $P_{LOSS}$ [W]</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.5</td>
<td>0.2</td>
<td>0.5</td>
</tr>
<tr>
<td>1.0</td>
<td>0.5</td>
<td>0.8</td>
</tr>
<tr>
<td>1.5</td>
<td>0.7</td>
<td>1.0</td>
</tr>
<tr>
<td>2.0</td>
<td>0.9</td>
<td>1.5</td>
</tr>
<tr>
<td>2.5</td>
<td>1.2</td>
<td>2.0</td>
</tr>
<tr>
<td>3.0</td>
<td>1.5</td>
<td>3.0</td>
</tr>
<tr>
<td>3.5</td>
<td>2.0</td>
<td>4.0</td>
</tr>
</tbody>
</table>

---

**Equations:**

1. $P_{CPZ} = 2 \left( \frac{E_{XFR} - E_{PC}}{t_{VIB}} \right)$
2. $P_{LOSS} = P_{RPZ} + P_{CPZ} + P_{RL} + P_{MR} + P_{MG} + P_{ST} + P_{Lk} + P_{BL}$
is higher. The switched inductor here raises \( v_{PZ} \) by pre-charging \( C_{PZ} \) to \( v_{PC} \) before every half cycle begins. This way, \( i_{PZ} \) charges \( C_{PZ} \) to a higher peak \( v_{PZ(PK)} \) to \( v_{PC} + \Delta v_{PZ(OC)} \) (not just \( \Delta v_{PZ(OC)} \)). But since \( C_{PZ} \) exposes the switches in the bridge to this \( v_{PZ(PK)} \), \( v_{PZ(PK)} \) cannot exceed the transistors’ breakdown level \( v_{BD} \):

\[
\v_{PZ(PK)} = v_{PC} + \Delta v_{PZ(OC)} \leq v_{BD}.
\]

(7)

\( v_{PZ(PK)} \) does not exceed \( v_{BD} \) when losses limit and set maximum output power \( P_{O(MPP)} \), like in Figs. 10 and 11 when \( i_{PZ(PK)} \) is below 21 \( \mu \)A. Above 21 \( \mu \)A, however, additional losses \( P_{LOSS} \) balance gains in piezoelectric power \( P_{PZ} \) at levels that require \( v_{PZ(PK)} \) to surpass \( v_{BD} \). \( v_{BD} \) limits \( v_{PZ(PK)} \) under those conditions, which then limits and sets \( P_{O(MPP)} \) before \( P_{LOSS} \) does. So as stronger vibrations raise \( \Delta v \) above 3.5 V in Figs. 11 and 12, the system reduces \( C_{PZ} \)’s pre-charge level \( v_{PC} \) to keep \( v_{PZ(PK)} \) from exceeding \( v_{BD} \)’s 5.5-V limit. This is why \( C_{PZ} \)’s loss \( P_{CPZ} \) falls in this region, because \( v_{PC} \) drops when \( \Delta v_{PZ(OC)} \) surpasses 5.5 V, and \( v_{BD} \) keeps \( v_{PC} \) so low that \( C_{PZ} \)’s energy \( E_{PK} \) is too much for \( L_X \) to pre-charge \( C_{PZ} \) to \( v_{PC} \). When \( \Delta v_{PZ(OC)} \) surpasses 4.0 V, however, \( v_{BD} \) keeps \( v_{PC} \) so low that \( C_{PZ} \)’s energy \( E_{PK} \) is too much for \( L_X \) to pre-charge \( C_{PZ} \) to \( v_{PC} \). So before \( C_{PZ} \) finishes draining into \( L_X \), the system steers some of this energy \( E_{PK} \) to \( v_{B} \). As a result, \( L_X \) receives and transfers less energy, and with less current to conduct, series resistances burn less ohmic power \( P_R \). Since fractional losses \( P_{LOSS}/P_{PZ} \) are lower, power-conversion efficiency \( \eta_c \) increases in this region.

IV. MEASURED PROTOTYPE

The CMOS die in Fig. 14 integrates the power switches, the drivers, and the maximum-supply selector in Figs. 2 and 7. The board shown incorporates the packaged integrated circuit (IC) and the 330-\( \mu \)H, 1.6-\( \Omega \) transfer inductor \( L_X \) used. The 16.8-nF, 9.5-\( \Omega \) piezoelectric cantilever transducer attaches to the edge of the board. The die, transducer, and inductor occupy 0.7 \( \times \) 0.5 mm\(^2\), 10 \( \times \) 50 \( \times \) 1 mm\(^3\), and 5 \( \times \) 5 \( \times \) 1 mm\(^3\), respectively. A field-programmable gate array (FPGA) generates the signals that switch the switched-inductor bridge and a shaker from Brüel & Kjær vibrates the cantilever so it generates 3–33 \( \mu \)A.

A. Power Indices

Maximum output power \( P_{O(MPP)} \) hinges on vibration frequency and strength, the transducer, the voltage with which the transducer’s current generates power, and losses. Of these, the only design-independent, application-specific factors are vibration frequency \( f_{VIB} \) and strength (in the form of acceleration in the mechanical domain and current \( i_{PZ} \) in the electrical domain). The transducer and the circuit that transfers power are typically the design variables that engineers control. Unfortunately, research splits along these lines, so advancements in circuits normally appear in the absence of advancements in transducers, and vice versa. As a result, literature rarely reports the best all-around solution.

The research focus here and in [13]–[19], [30]–[35] is on...
the circuit, not the transducer. So comparing performance without normalizing the effects of the transducers used seems unfair. The components in the transducer that in part determine \( P_{\text{OMPP}} \) are capacitance \( C_{\text{PZ}} \), resistance \( R_{\text{PZ}} \), and dynamic capacitance \( C_{\text{PZ(XFR)}} \). But since the diode bridges in [13]-[15], [19] draw power across half cycles (not between half cycles), the dynamic effects of \( C_{\text{PZ(XFR)}} \) are absent. So including \( C_{\text{PZ(XFR)}} \)'s losses \( P_{\text{PZ}} \) in the comparison seems fair. How \( R_{\text{PZ}} \) affects \( P_{\text{O(MPP)}} \) also depends on the harvester, so additional power a harvester draws (with \( P_{\text{PZ}} \)) than a lossless case is a good way of assessing and normalizing circuit performance to the transducers used.

In this light, drawn power index \( \eta_{\text{PZ}} \) indicates how much additional power a harvester draws (with \( P_{\text{PZ}} \)) than a lossless diode bridge can harness from the same piezoelectric transducer:

\[
\eta_{\text{PZ}} = \frac{P_{\text{PZ}}}{P_{\text{PZ(BRDG)}}}. \tag{10}
\]

Losses \( P_{\text{LOSS}} \) limit how much of \( P_{\text{PZ}} \) the system ultimately delivers with \( P_{\text{OMPP}} \). So with respect to the output, power-conversion efficiency \( \eta_{\text{c}} \) reduces \( \eta_{\text{PZ}} \) to yield output power index \( \eta_{\text{O(MPP)}} \):

\[
\eta_{\text{O(MPP)}} = \eta_{\text{PZ}} \eta_{\text{c}} = \left( \frac{P_{\text{PZ}}}{P_{\text{PZ(BRDG)}}} \right) \left( \frac{P_{\text{OMPP}}}{P_{\text{PZ}}} \right) = \frac{P_{\text{OMPP}}}{P_{\text{PZ(BRDG)}}}, \tag{11}
\]

which compares \( P_{\text{O(MPP)}} \) with \( P_{\text{PZ(BRDG)}} \). But since gains in \( P_{\text{PZ}} \) can outpace \( P_{\text{LOSS}} \) and vice versa, \( \eta_{\text{O(MPP)}} \) is a better metric than either \( \eta_{\text{PZ}} \) or \( \eta_{\text{c}} \) alone. Still, \( \eta_{\text{PZ}} \) indicates the ability of a system to draw piezoelectric power \( P_{\text{PZ}} \) and \( \eta_{\text{c}} \) the efficacy with which a system transfers this power.

### B. Performance

The 3.3–33 \( \mu \text{A} \) (\( i_{\text{PZ}} \)) that the shaker here induces at 120 Hz (\( f_{\text{VIB}} \)) charges a 16.8-nF piezoelectric capacitance \( C_{\text{PZ}} \) across 0.5–5.5 V (\( \Delta V_{\text{PZ(OCC)}} \)). With this stimulation, the lossless diode bridge can draw 0.13–15 \( \mu \text{W} \), like \( P_{\text{PZ(BRDG)}} \) in Fig. 15 shows. Under the same conditions, the series switched-inductor bridge prototyped here draws 1.2–55 \( \mu \text{W} \) (\( P_{\text{PZ}} \)) and delivers 0.70–49 \( \mu \text{W} \) (\( P_{\text{OMPP}} \)). So like Fig. 15 shows, the system draws 3.7×–9.8× (\( \eta_{\text{PZ}} \)) and outputs 3.3×–6.8× (\( \eta_{\text{O(MPP)}} \)) more power than the lossless bridge can and outputs 45%–91% (\( \eta_{\text{c}} \)) of the power drawn.

**Fig. 15.** Measured power and resulting power indices.

Piezoelectric power \( P_{\text{PZ}} \) overcomes losses when vibrations charge \( C_{\text{PZ}} \) more than 0.5 V. This 0.5-V threshold corresponds to the minimum vibration strength from which the system can harvest power. Power indices \( \eta_{\text{PZ}} \) and \( \eta_{\text{O(MPP)}} \) are high at 5.9–9.8 and 4.5–6.8 when vibrations charge \( C_{\text{PZ}} \) across 0.5–3.5 V.

#### TABLE II: RELATIVE PERFORMANCE

<table>
<thead>
<tr>
<th>Diode Bridge</th>
<th>Half-Switched SL Bridge</th>
<th>SL Half Bridge</th>
<th>Switched-Inductor Recycling Bridges</th>
<th>SL Bridge</th>
</tr>
</thead>
<tbody>
<tr>
<td>( L_{\text{MIN}} )</td>
<td>1 ( \mu \text{m} )</td>
<td>2 ( \mu \text{m} )</td>
<td>180 nm</td>
<td>250 nm</td>
</tr>
<tr>
<td>( V_{\text{bb}} )</td>
<td>15 V</td>
<td></td>
<td></td>
<td>5 V</td>
</tr>
<tr>
<td>Si Area</td>
<td>4.25 mm²</td>
<td>0.90 mm²</td>
<td>2.3 mm²</td>
<td>0.75 mm²</td>
</tr>
<tr>
<td>( L_{\text{x}} )</td>
<td>160 ( \mu \text{H} )</td>
<td>330 ( \mu \text{H} )</td>
<td>220 ( \mu \text{H} )</td>
<td>1000 ( \mu \text{H} )</td>
</tr>
<tr>
<td>( R_{\text{L}} )</td>
<td>3.4 ( \Omega )</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>( \Delta V_{\text{PZ(OC)}} )</td>
<td>40.7 ( \mu \text{A} )</td>
<td>30.2–104 ( \mu \text{A} )</td>
<td>8.22–35.7 ( \mu \text{A} )</td>
<td>84 ( \mu \text{A} )</td>
</tr>
<tr>
<td>( C_{\text{PZ}} )</td>
<td>12 nF</td>
<td>275 nF</td>
<td>15 nF</td>
<td>19 nF</td>
</tr>
<tr>
<td>( f_{\text{VIB}} )</td>
<td>225 Hz</td>
<td>100 Hz</td>
<td></td>
<td>143 Hz</td>
</tr>
<tr>
<td>( \Delta V_{\text{PZ(OCC)}} )</td>
<td>4.8 V</td>
<td>0.35–1.2 V</td>
<td>1.22–5.24 V</td>
<td>9.8 V</td>
</tr>
<tr>
<td>( P_{\text{PZ(BRDG)}} )</td>
<td>15.6 ( \mu \text{W} )</td>
<td>0.84–9.9 ( \mu \text{W} )</td>
<td>0.80–15.1 ( \mu \text{W} )</td>
<td>65.7 ( \mu \text{W} )</td>
</tr>
<tr>
<td>( P_{\text{PZ}} )</td>
<td>8.2 ( \mu \text{W} )</td>
<td>4–72 ( \mu \text{W} )</td>
<td>7–78 ( \mu \text{W} )</td>
<td></td>
</tr>
<tr>
<td>( \eta_{\text{PZ}} )</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>( \eta_{\text{O(MPP)}} )</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Component</td>
<td>4 FETs MPP Buffer</td>
<td>L(_x), 4 FETs</td>
<td>L(_x), 6 FETs MPP Buffer</td>
<td>L(_x), 6 FETs MPP Buffer</td>
</tr>
</tbody>
</table>

\( ^{a} \text{Calculated from } C_{\text{PZ}}'s \text{ open-circuit voltage variation } \Delta V_{\text{PZ(OCC)}}. \)\( ^{b} \text{Increased to discount controller's } P_{\text{Q}}. \)\( ^{c} \text{Approximated from reported graphs.} \)
Harvesting performance is lower when vibrations are stronger because the transistors’ 5.5-V breakdown voltage $V_{BD}$ limits the voltage $V_{PZ}$ with which $i_{PZ}$ draws $P_{PZ}$.

Power-conversion efficiency $\eta_C$ is not as important as $\eta_{iPZ}$ and $\eta_{(iO)}$ because $\eta_C$ does not account for how much power the harvester can draw from motion. $\eta_C$ is still a good measure of design quality, however, because systems output more power when optimally designed to balance losses. In this case, $\eta_C$ is $75\%$–$91\%$ when vibrations charge $C_{PZ}$ more than 3 V. $\eta_C$ falls to 45% below 1 V because gate-drive and shoot-through losses $P_{G}$ and $P_{ST}$, which do not scale down with $P_{PZ}$, dominate when $P_{PZ}$ is low. In other words, the system is more optimal for higher $P_{PZ}$.

C. Relative Performance

Diode bridges output less power than switched-inductor (SL) bridges because the voltage $V_{PZ}$ with which piezoelectric current $i_{PZ}$ generates power $P_{PZ}$ is lower in diode bridges than the voltage $\Delta V_{PZ(OC)}$ that $i_{PZ}$ charges $C_{PZ}$ across a half cycle in switched inductors [37]. Plus, and this is also why $V_{PZ}$ never reaches $\Delta V_{PZ(OC)}$ in diode bridges, diode bridges collect a fraction of the charge that $i_{PZ}$ outputs. The diode bridge [14] in Table II, for example, draws 53% and outputs 48% of the power that the lossless diode bridge can transfer. Table II summarizes steady-state performance across vibration cycles. Since the controller here is off chip, parameters $P_{O(MPP)}$, $\eta_C$, and $\eta_{(iO)}$ in the table discount the power consumed by the controller. This way, the table compares the performance of the power stage only (without the controller).

The half-switched switched-inductor diode bridge in [15] draws $4.8 \times 7.3 \times$ and outputs $1.2 \times 3.0 \times$ more power than the lossless diode bridge can because the system collects all the charge that $i_{PZ}$ outputs with a higher voltage (that is as high as $\Delta V_{PZ(OC)}$). The switched-inductor half bridge in [16] draws and outputs more power ($5.4 \times 11 \times$ and $2.6 \times 3.5 \times$) because the system charges $C_{PZ}$ above $\Delta V_{PZ(OC)}$. The recycling switched-inductor bridges in [17]–[19], [38]–[41] draw and output even more power because $V_{PZ}$ is, on average, greater than the others. The redeeming advantage of the diode bridge is that it does not require a controller, so quiescent losses are absent.

The switched-inductor bridge prototyped here outputs up to $6.8 \times$ more power than the lossless bridge can. This is as much power as the recycling bridge in [19] outputs, but with fewer power switches. In addition to the six switches that the rectifier uses, [19] also requires a two-switch bucking power stage to keep the rectified level at the maximum power point.

Reported silicon areas are not comparable because the minimum channel length $L_{MIN}$ for the 5-V 0.54-mm2 switches in [19] is 350 nm and $L_{MIN}$ for the 5.5-V 0.25-mm2 transistors here is 1.2 μm for NFETs and 810 nm for PFETs. Plus, the switches in [19] transfer 410 μW and the transistors here transfer 49 μW. Still, four fewer power switches without a separate MPP power stage under similar constraints should consume less power, and as a result, deliver more of the power drawn. This is probably why the minimum harvestable voltage $\Delta V_{PZ(OC)}$ here is 840 mV lower at 500 mV than the 1.34 V needed in [19].

V. CONCLUSIONS

The switched-inductor bridge presented here draws 1.2–55 μW and outputs 0.70–49 μW with the 3.3–33 μA that a vibrating 16.8-nF piezoelectric transducer generates at 120 Hz to charge the transducer across 0.5–5.5 V. The prototype draws $3.7 \times 9.8 \times$ and outputs $3.3 \times 6.8 \times$ more power than a lossless diode bridge can. This is as much power as the best recycling switched-inductor bridge outputs, but with four fewer power switches and without a separate maximum power-point stage. This is possible because, before every half cycle begins, the single-inductor four-switch network pre-charges the transducer to an adjustable voltage. This way, the system can deliver as much power as power losses and breakdown voltage allow. Drawing maximum power is important because small piezoelectric transducers couple a very small fraction of the mechanical power that vibrations supply. Replenishing the energy that drains tiny onboard batteries with higher power extends the cost-, energy-, and life-saving features that biomedical implants and wireless microsensors networked across fields, factories, hospitals, and homes can offer.

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REFERENCES

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