Tiny Piezoelectric Harvesters: Principles, Constraints, and Power Conversion

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Abstract—Wireless microsystems can add intelligence to hospitals, homes, and factories that can save money, energy, and lives. Unfortunately, tiny batteries cannot store sufficient energy to sustain useful microsystems for long, and replacing or recharging the batteries of hundreds of networked nodes is costly and invasive in the case of the human body. Thankfully, shocks and vibrations are prevalent in many applications, so ambient kinetic energy can continually replenish batteries to extend the life of the systems they support. And since tiny devices produce minimal damping effects on motion, they can draw as much power as the microelectronics allow. Unfortunately, uncollected charge, breakdown voltages, and energy losses limit how much power harvesting microsystems can generate. This is why this paper reviews how tiny transducers generate power and how state-of-the-art diode bridges and switched inductors and their derivatives draw and output as much power as possible. Of prevailing technologies, in fact, the recycling bridge pre-damps the transducer at the highest voltage possible all the time to output the highest power. But because it still needs a regulating charger to stay at its maximum power point, other pre-damping switched inductors suffer lower losses and require less space. Although the pre-damping bridgeless solution pre-damps every other half cycle, it generates comparable power with only two switches. No harvester, however, escapes the limits that power losses and breakdown voltages impose, so output power is always finite, and in the case of miniaturized systems, not very high.

Index Terms—Piezoelectric harvesters, ambient kinetic energy, motion, vibration, shock, diode bridges, switched inductors, harvesting chargers, and energy-harnessing microsystems.

I. HARVESTING MICROSYSTEMS

Wireless microsystems that sense, monitor, manage, and report information in hospitals, homes, office buildings, factories, farms, and humans can save money, energy, and lives [1]. Although the power they require is nowadays low in the microwatt range [2], miniaturized batteries cannot store sufficient energy to sustain them for months or years at a time [3]. This is why drawing power from ambient sources is so popular today, because the environment is a vast tank that requires no board space.

Photovoltaic cells are popular in this respect because sunlight outputs the highest power levels at 10–15 mW/cm² [4]. The problem is solar light is often unavailable, and artificial lighting produces less than 100 µW/cm² [5].

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Thankfully, shocks and vibrations are prevalent in automobiles, airplanes, machinery, and humans. Plus, electromechanical transducers can output moderate power. Of these, though, piezoelectric devices can produce more power at 300 µW/cm² than their electrostatic and electromagnetic counterparts can at 100 µW/cm² and 10 µW/cm² [6].

Still, micro-scale piezoelectric transducers harness a small fraction of ambient kinetic power, and only in the presence of shocks and vibrations. A system must therefore incorporate a rechargeable battery vBAT that can supply power when ambient power is insufficient or unavailable [7]. The role of the transducer in Fig. 1, for example, is to recharge vBAT. The aims of the piezoelectric harvester [8], feedback loop, and power supply are to draw power, keep the system at its maximum power point (MPP), and feed the power amplifier (PA), digital-signal processor (DSP), analog–digital converter (ADC), and the other functional blocks that comprise the load.

Fig. 1. Piezoelectric energy-harvesting microsystem.

This paper surveys, compares, and derives from the state of the art how tiny piezoelectric harvesters can generate power. Although not to the same extent, breadth, or depth, [8] and [40] similarly review some (but not all) the technologies and concepts presented here. In this rendition, Section II reviews how and under which conditions state-of-the-art micro-scale transducers output charge in the presence of motion. Sections III, IV, and V then describe and assess the latest in piezoelectric chargers and their derivatives. Sections VI and VII end with comparisons and conclusions.

II. TINY PIEZOELECTRIC TRANSDUCERS

A. Basic Operation

When unstrained, piezoelectric material is electrically neutral because, as Fig. 2a illustrates, positive and negative charge centers align and balance. The atomic arrangement is such, however, that charge centers shift away from one another when the material strains [9]. This shift produces a surface potential that changes continuously with variations in mechanical deformation.

A transducer must therefore incorporate a piece of piezoelectric material that strains in the presence of shocks...
and vibrations. For this, only one end of the piezoelectric cantilever in Fig. 2b, for example, is stationary on a firm base and the other carries a mass [10]. This way, motion produces a displacement and strain that the mass extends. Once maximally strained, the cantilever springs back, and under favorable conditions, oscillates. The frequency at which the device will most easily oscillate depends on the tensile and elastic properties of the material and the size and weight of the mass. Although other mechanical configurations are possible, similar principles govern their design, operation, and properties.

![Fig. 2. Piezoelectric (a) material and (b) transducer.](image)

Table I lists the state of the art in tiny piezoelectric transducers [10], [12]–[15]. Unpackaged volumes range from 12 to 27 mm³. They output 10–85 µW at 100–300 µW/cm², which is higher than what electrostatic transducers usually can [6], but as expected, lower than what sunlight can [4]. Natural frequencies range from 252 to 1300 Hz, which is slightly higher than the 1–300 Hz vibrations that most applications exhibit [11]. Unfortunately, [10], [12]–[15] did not quote capacitance, so how they compare with the nF's that commercially available transducers exhibit is not certain.

<table>
<thead>
<tr>
<th>Reference</th>
<th>Unpackaged Volume</th>
<th>Power</th>
<th>Frequency</th>
</tr>
</thead>
<tbody>
<tr>
<td>[12]</td>
<td>12 mm³</td>
<td>10 W</td>
<td>255 Hz</td>
</tr>
<tr>
<td>[10]</td>
<td>19 mm³</td>
<td>35 µW</td>
<td>572 Hz</td>
</tr>
<tr>
<td>[13]</td>
<td>0.021 mm³</td>
<td>45 µW</td>
<td>1300 Hz</td>
</tr>
<tr>
<td>[14]</td>
<td>27 mm³</td>
<td>68 µW</td>
<td>419 Hz</td>
</tr>
<tr>
<td>[15]</td>
<td>Not Reported</td>
<td>85 µW</td>
<td>325 Hz</td>
</tr>
</tbody>
</table>

Volume of the active piezoelectric material (not the transducer).

![Image](image)

**B. Electrical Model**

When motion pushes the cantilever away from its resting place, the material acquires a potential energy $E_{PE}$ that climbs with displacement $d_s$ and peaks when maximally displaced. The spring then pushes the mass $M_S$ back with such force that velocity $v_S$ and kinetic energy $E_{KE}$ peak when $d_s$ is back at zero. So mass swings in the other direction to repeat the process. The system continually exchanges potential and kinetic energies this way until impeding forces damp oscillations.

Like a spring, an inductor $L_M$ can receive or release energy, but not hold it. So with a capacitor $C_M$ in parallel, $L_M$'s energy $E_L$ first drains to $C_M$ and $C_M$'s energy $E_C$ then depletes back into $E_L$ and alternate this way until parasitic resistances $R_M$ burn the energy. Since $E_L$ and $E_C$ rise with $L_M$'s current $i_{L_M}$ and $C_M$'s voltage $v_{CM}$ like $E_{PE}$ and $E_{KE}$ with displacement $d_S$ and velocity $v_S$, $i_{LM}$ and $v_{CM}$ in Fig. 3 can mimic $d_S$ and $v_S$ when $L_M$ maps to the cantilever's spring constant $K_S$, $C_M$ to $M_S$, and $R_M$ to mechanical damping forces:

$$E_{PE} = 0.5K_Sv_S^2 = E_L = 0.5L_Mi_{LM}^2 \quad (1)$$  

A Norton-equivalent current source $i_S$ can therefore emulate the shocks and vibrations that power the transducer. Thevenin-equivalent models, which use voltage sources and series components, can also model this behavior [16].

![Image](image)

The purpose of the transducer is to map mechanical energy into the electrical domain. In this case, $M_S$'s kinetic energy in $C_M$ corresponds to electrostatic energy in the cantilever's capacitance $C_PZ$. But since velocity $v_{CM}$ does not equate to $C_PZ$ voltage, the transformer turns ratio $N_{PZ/M}$ maps $v_{CM}$ to $v_{PZ}$:

$$N_{PZ} = \frac{v_{KL}}{v_{CM}}. \quad (3)$$

$R_{PZ}$ then models the power that $C_PZ$ leaks.

Power derived from the transducer is a power drain that damps oscillations. Of the power drawn with $P_E$ in the mechanical domain, however, only a fraction couples to the electrical domain with $P_{Z}$. This is why the power-converting transformer in Fig. 3 delivers an electromechanical fraction $k_c^2$ of what $P_E$ supplies to $P_{PZ}$ and keeps the rest in the mechanical domain. So $P_E$'s equivalent resistance $R_E$ in Fig. 4 effectively burns $1/k_c^2$ more power than $P_{PZ}$'s equivalent load $R_L$ or $v_{PZ}/i_{PZ}$:

$$R_E = \frac{v_{CM}^2}{P_E} = \frac{k_c^2}{k_{PZ}} \frac{v_{CM}^2}{P_{CM}} = \frac{k_c^2}{v_{PZ}} = \frac{R_L}{N_{CM}} \left(\frac{N_{PZ}}{N_{CM}}\right)^2. \quad (4)$$

![Image](image)

**C. Maximum Power**

When vibration frequencies matches the resonant frequency of the tank, motion feeds energy that $L_M$ and $C_M$ exchange to produce an oscillating voltage that is in phase with $i_S$. As a result, $i_S$ delivers power $P_S$ across both positive and negative half cycles. The energy in the tank therefore grows as long as $P_S$ surpasses $R_M$ and $R_E$'s combined losses. When losses match $P_S$, $L_M$ and $C_M$ supply each other's needs and $R_M$ and $R_E$ consume $P_S$. In other words, $L_M$ and $C_M$ in Fig. 4, for all
practical purposes, disappear and $R_M$ and $R_E$ receive all of $P_S$.

When vibration and resonant frequencies do not match, however, $V_M$ is no longer in phase with $i_S$. Although $i_S$ still produces power when $i_S$ and $V_M$ are both positive and both negative, $i_S$ now consumes power when their polarities oppose. The transducer therefore draws the greatest power from motion when tuned to vibrations.

Since additional $P_S$ has little effect on the velocity and displacement of tiny under-damped devices, the fundamental aim of a harvester should be to draw the highest power possible. For this, the circuit should keep $V_{PZ}$ in phase with $i_{PZ}$ and at the highest level that the breakdown voltage allows. Except, conduction losses climb with drawn power, so when still below the breakdown level, $V_{PZ}$ should rise until incremental losses cancel additional gains. In other words, frequency, breakdown voltage, and losses limit output power.

III. BASIC DIODE BRIDGES

The basic diode bridge in Fig. 5 is popular in this space because the diodes rectify and steer $i_{PZ}$ into a receiving capacitor $C_{REC}$ [19]. When assuming diode voltages are zero, for example, $i_{PZ}$’s positive half cycle in Fig. 6 charges $C_{PZ}$ until $V_{PZ}$ overcomes $C_{REC}$’s rectified output voltage $V_{REC}$. Past that point and through the end of the positive half cycle, diodes $D_{PG}$ and $D_{PG}$ steer $i_{PZ}$ into $C_{REC}$. Similarly, $i_{PZ}$ across negative half cycles discharges $C_{PZ}$ until $D_{NG}$ and $D_{NO}$ clamp $V_{PZ}$ to $-V_{REC}$, past which point $i_{PZ}$ flows, again, into $C_{REC}$.

To quantify this point, consider that, without the bridge, $i_{PZ}$’s half-cycle charge $q_{HALF}$ in Fig. 6 would charge $C_{PZ}$ across peak–peak open-circuit voltage $\Delta v_{PZ(OC)}$, so

$$q_{HALF} = C_{PZ} \Delta v_{PZ(OC)}. \tag{5}$$

But since $C_{PZ}$ absorbs some of $q_{HALF}$ when charging across $2V_{REC}$, $C_{REC}$ loses charge $q_{LOST}$ to $C_{PZ}$:

$$q_{LOST} = C_{PZ} (2V_{REC}). \tag{6}$$

$$C_{REC}$$ therefore collects with $q_{REC}$ the difference every half cycle to harness with $E_{H}$ twice $q_{REC}$’s energy per cycle:

$$E_{H} = 2(q_{HALF} - q_{LOST})V_{REC} = 2C_{PZ} (\Delta v_{PZ(OC)} V_{REC} - 2V_{REC}^2). \tag{7}$$

The maximum power point results when the incremental loss in $q_{LOST}$ balances the additional gain in $q_{HALF}$, which happens when the combined derivative is zero and $V_{REC}$ is $0.25\Delta v_{PZ(OC)}$:

$$\frac{dE_{H}}{dv_{REC}} = 2C_{PZ} (\Delta v_{PZ(OC)} - 4V_{REC} |\Delta v_{PZ(OC)} = 0.25V_{REC}/2| = 0. \tag{8}$$

In other words, $E_{H}$ peaks to $0.25C_{PZ}\Delta v_{PZ(OC)}^2$ when the loaded swing $\Delta v_{PZ(LD)}$ is half the open-circuit counterpart $\Delta v_{PZ(OC)}$:

$$E_{HMAX}^2 = 0.25C_{PZ} \Delta v_{PZ(OC)}^2. \tag{9}$$

Output power $P_O$ can therefore be $0.25C_{PZ}\Delta v_{PZ(OC)}^2 \eta_0$.

4. Cross-Coupled Bridge

Unfortunately, actual diodes drop 0.6–0.8 V when they conduct $i_{PZ}$, so they burn substantial power. Luckily, $V_{REC}$ is sometimes greater than an NMOS threshold voltage $V_{TN}$. As a result, $V_{PZ}$ rises high enough to close the cross-coupled NFETs in Fig. 7. Or viewed differently, $V_{PZ}$ falls low enough below $V_{REC}$ to close a cross-coupled PFET pair [20] in place of $D_{PG}$ and $D_{NO}$ in Fig. 5. This way, $M_{NG}$ and $M_{PG}$ drop less than 100 mV to reduce conduction losses by more than 83%.

![Fig. 5. Diode-bridge charger.](image)

![Fig. 6. Ideal diode-bridge waveforms.](image)

Interestingly, raising $V_{REC}$ increases the power $i_{PZ}$ delivers into $C_{REC}$ at $V_{REC}$ as well as the charge lost to $C_{PZ}$ when swinging $V_{PZ}$ between $V_{REC}$ and $-V_{REC}$. So for maximum power, the system should raise $V_{REC}$ until the incremental loss cancels the additional gain, which happens at a particular $V_{REC}$. This is why a charger in Fig. 5 draws just enough power from $C_{REC}$ to keep $V_{REC}$ near its maximum power point.

Notice $M_{NG}$ and $M_{PG}$ close whenever $V_{PZ}$ rises above $V_{TN}$, which happens before $V_{PZ}$ reaches $V_{REC}$. This is acceptable because $D_{PG}$ and $D_{NO}$ block $i_{PZ}$ until $V_{PZ}$ reaches $V_{REC}$. This is why PFETs cannot replace $D_{PG}$ and $D_{NO}$ when NFETs already replace $D_{NG}$ and $D_{PG}$ because PFETs would also engage early. Shorting $C_{PZ}$ to $C_{REC}$ this way when $V_{PZ}$ is below $V_{REC}$ would drain $C_{REC}$ into $C_{PZ}$. Replacing $D_{PG}$ and $D_{NO}$ with cross-coupled PFETs and inserting one diode $D_{REC}$ between them and $C_{REC}$, however, like Fig. 8 [21] shows, works because $D_{REC}$ blocks $i_{PZ}$. The tradeoff for reducing two diodes to one is adding one series resistance to the conduction path.

![Fig. 7. Cross-coupled diode-bridge charger.](image)

![Fig. 8. Cross-coupled diode-bridge charger (negative voltage converter NVC).](image)
B. Half Bridge

The half bridge in Fig. 9 also rectifies and steers \(i_{PZ} \) into \(C_{REC} \) [22]–[23]. Similar to the ideal full bridge, \(i_{PZ} \)'s positive half-cycle charge in Fig. 10 charges \(C_{PZ} \) until \(v_{PZ} \) overcomes \(v_{REC} \), past which point \(D_{REC} \) conducts \(i_{PZ} \) into \(C_{REC} \). \(i_{PZ} \)'s negative half-cycle charge then drains \(C_{PZ} \) until \(D_{G} \) clamps \(v_{PZ} \) to 0 V and steers the rest of \(i_{PZ} \)'s negative half-cycle charge to ground. This means, \(C_{REC} \) only harnesses positive charge.

\[ \text{Fig. 9. Half-bridge charger.} \]

![Image](image.png)

![Image](image.png)

**Fig. 10. Ideal half-bridge waveforms.**

Ultimately, \(C_{REC} \) collects the positive charge that \(i_{PZ} \) does not lose to \(C_{PZ} \) when swinging \(v_{PZ} \) across \(v_{REC} \). But like before, \(i_{PZ} \) delivers more power to \(C_{REC} \) with a higher \(v_{REC} \) and \(C_{PZ} \) loses more charge when swinging across a wider \(v_{REC} \). So maximum power results at the \(v_{REC} \) that balances this tradeoff. This is why Fig. 9 includes a charger, to draw just enough power from \(C_{REC} \) to keep \(v_{REC} \) near its maximum power point.

To derive this maximum power point, first recall that \(i_{PZ} \)'s positive charge \(q_{PZ} \) is still \(C_{REC}v_{PZ} \). But since \(i_{PZ} \) loses positive charge \(q_{PZ} \) when charging to \(v_{REC} \):

\[ q_{PZ} = C_{REC}v_{REC}, \]

\[ C_{REC} \text{ collects with } q_{REC} \text{ and } E_{H} \text{ the difference at} \]

\[ E_{H} = (q_{PZ} - q_{LOST})v_{REC} = C_{REC}\Delta v_{PZ}, \]

\[ \Delta v_{PZ} = \frac{1}{2}C_{REC} \frac{dE_{H}}{dv_{REC}} = \frac{1}{2}C_{REC}C_{REC} = \frac{1}{2}v_{REC}. \]

Maximum power results when the incremental loss cancels the additional gain, or when the combined derivative is zero and \(v_{REC} \) is twice that of the full bridge at 0.5\( \Delta v_{PZ} \): \(v_{REC} \) is how saved conduction power stacks against quiescent power and the power lost because of the comparator's delay. In other words, if saved power does not outweigh losses, which can be the case when \(i_{PZ} \) is low, a diode can be a lower-loss option.

C. Diode Options

As already mentioned, diodes can burn substantial conduction power when they steer \(i_{PZ} \) because they drop 0.6–0.8 V. Although replacing two diodes in the full bridge with cross-coupled FETs is possible when \(v_{REC} \) is greater than a MOS threshold voltage, the same is not true for all diodes in the full bridge or diodes in the half bridge. And if \(v_{REC} \) is less than 0.6–0.8 V, like in Fig. 6, replacing two diodes is not even possible.

Ideally, a diode should drop 0 V, lose 0 A, and respond instantly. Although on-chip Schottky and P–N junction diodes drop 0.4–0.6 V and 0.6–0.8 V, they lose no ground current and respond almost instantly. Similarly, the diode-connected FET in Fig. 11b drops a gate–source voltage \(v_{GS} \) that can be 0.6–0.9 V, loses 0 A, and although not to the same extent, responds very quickly.

\[ \text{Fig. 11. Diode options.} \]

The voltage source \(v_{S} \) in Fig. 11c shifts the FET's \(v_{GS} \) by \(v_{S} \) so the switch can drop \(v_{GS} - v_{S} \), which can be 100–200 mV [24] and correspond to a 70%–90% reduction in conduction power. The voltage is not lower because the tolerance of MOS threshold voltages is high on the order of ±75–±100 mV [25], so margin must exist to ensure the FET does not conduct reverse current. Irrespective, the tradeoffs for this reduction in conduction power are quiescent power and response time, because \(v_{S} \) is a circuit that requires power and time to react.

The FET in Fig. 11d drops an even lower voltage because the comparator can overdrive the FET into triode when terminal voltages are only millivolts apart [26]. This way, the drain–source voltage \(v_{DS} \) can be 25–100 mV and the corresponding conduction loss 80%–97% lower than a diode. Like before, though, the tradeoffs are quiescent power and response time. But since conduction losses and tolerance are lower, this option is often preferable over the shifted counterpart.

Adding an input offset to the comparator so it transitions early can offset the delay of the circuit to reduce its effective response time [27]. Too much offset, however, can trip the comparator before it should. A more important consideration is how saved conduction power stacks against quiescent power and the power lost because of the comparator's delay. In other words, if saved power does not outweigh losses, which can be the case when \(i_{PZ} \) is low, a diode can be a lower-loss option.

IV. Basic Switched Inductors

A. Energy Transfers

The fundamental aim of inductors in energy harvesters is to transfer energy. When connected across a battery \(v_{BAT} \), for example, an inductor \(L_{X} \) draws current and energy from \(v_{BAT} \). In this case, \(L_{X} \)'s energizing voltage \(v_{0} \) is \(v_{BAT} \), and since \(v_{BAT} \) is fairly constant, \(L_{X} \)'s current \(i_{L} \) rises linearly with time at the rate of \(v_{L}/L_{X} \) like Fig. 12a shows. So when \(L_{X} \) collects with \(i_{L} \), the desired amount of energy \(1/2L_{X}i_{L}^{2} \) from \(v_{BAT} \) across energizing time \(t_{0} \), the system disconnects \(L_{X} \) from \(v_{BAT} \).

Draining \(L_{X} \) into a battery is similar, but in reverse. Here, once \(L_{X} \) has energy in the form of \(i_{L} \), the system connects \(v_{BAT} \) across \(L_{X} \) such that \(L_{X} \)'s voltage is negative. With such a deenergizing voltage \(v_{D} \), \(i_{L} \) falls linearly with time to drain into
\( \text{v}_{\text{BAT}} \) at the rate of \( \text{v}_\text{p}/L_X \). When \( L_X \) depletes, which happens when \( i_c \) is zero, the system disconnects \( L_X \) from \( \text{v}_{\text{BAT}} \).

**Fig. 12.** Energy transfers with (a) batteries and (b) capacitors.

Drawing and supplying energy to a capacitor \( C_X \) is basically the same. When connecting an empty \( L_X \) across a charged \( C_X \), for example, \( C_X \) discharges into \( L_X \). But since \( C_X \)'s voltage falls as \( C_X \) drains, \( i_c \)'s rising rate decreases with time to produce the quarter sinusoid shown in Fig. 12b. Irrespective of \( C_X \)'s initial energy, \( C_X \) drains completely into \( L_X \) after a quarter resonance period 0.25\( t_{\text{LC}} \). Supplying a capacitor is the same, but in reverse. When connecting an energized \( L_X \) across an empty \( C_X \), \( L_X \)'s \( i_c \) charges \( C_X \), \( i_c \) drops more quickly as \( C_X \)'s voltage rises, and \( L_X \) fully depletes after 0.25\( t_{\text{LC}} \).

Since peak currents i\( \text{L}_{\text{PK}} \) inductances \( L_X \) and piezoelectric capacitances \( C_{\text{PZ}} \) in small commercially available transducers are usually below 100 mA, 300 \( \mu \)H, and 50 nF and battery voltages \( \text{v}_{\text{BAT}} \) are above 1 V [27], transfers complete within 4 \( \mu \)s. So of the 3 ms to 1 s that a typical cycle can last [11], each transfer normally requires less than 0.2% of the vibration period. This means, transfers are practically instantaneous.

To carry energy without consuming much power, inductor currents and resistances should be low. For low currents, inductances should be high, and for low resistances, coils should be large. This is why many implementations use 100-\( \mu \)H to 10-mH inductors that occupy more than 6 \times 6 \times 3 mm\(^3 \).

**B. Synchronized Switched-Inductor Discharges**

Basic switched-inductor harvesters use an inductor \( L_X \) to drain between cycles the charge that \( C_{\text{PZ}} \) collects across half cycles [29]. This way, with synchronous electric charge extraction (SECE), \( i_{\text{PZ}} \)'s positive half-cycle charge in Fig. 6 charges the unloaded \( C_{\text{PZ}} \) to \( \Delta \text{v}_{\text{PZ}(\text{OC})} \) in Fig. 13. At the end of \( i_{\text{PZ}} \)'s positive half cycle, \( L_X \) drains \( C_{\text{PZ}} \) into the battery \( \text{v}_{\text{BAT}} \). \( i_{\text{PZ}} \)'s negative half-cycle charge then charges \( C_{\text{PZ}} \) to \(-\Delta \text{v}_{\text{PZ}(\text{OC})}\) and \( L_X \) drains \( C_{\text{PZ}} \) into \( \text{v}_{\text{BAT}} \) at the end of the half cycle.

**Fig. 13.** Synchronized switched-inductor discharges.

Since \( \text{v}_{\text{BAT}} \) receives between half cycles the energy \( C_{\text{PZ}} \) collects across half cycles, \( \text{v}_{\text{BAT}} \) in one cycle harnesses \( E_H \) twice the energy \( C_{\text{PZ}} \) collects with \( \Delta \text{v}_{\text{PZ}(\text{OC})} \):

\[
E_H = E_{\text{C(+)}} + E_{\text{C(-)}} = 2 \left( 0.5 C_{\text{PZ}} \frac{\Delta \text{v}_{\text{PZ}(\text{OC})}}{2} \right) = C_{\text{PZ}} \Delta \text{v}_{\text{PZ}(\text{OC})}^2 .
\]  

Note this energy is 4\( \times \) higher than that of basic diode bridges. One reason for this is basic bridges lose \( i_{\text{PZ}} \) charge to \( C_{\text{PZ}} \) when swinging \( \text{v}_{\text{p}} \) between rectified limits. Another reason is \( C_{\text{PZ}} \)'s energy rises quadratically with \( \text{v}_{\text{PZ}} \), so the switched inductor collects more energy with 4\( \times \) the loaded swing at \( 2\Delta \text{v}_{\text{PZ}(\text{OC})} \) than the basic diode bridge does with 0.5\( \Delta \text{v}_{\text{PZ}(\text{OC})} \).

Although \( L_X \)'s series resistance (i.e., quality factor) and other losses limit some of these gains, a fundamental drawback and challenge with switched inductors is synchronizing switching events. Basic diode bridges draw piezoelectric power automatically whenever \( \text{v}_{\text{PZ}} \) overcomes its rectified output \( \text{v}_{\text{REC}} \). Switched inductors, on the other hand, must synchronize energy transfers to \( i_{\text{PZ}} \)'s half-cycle points. This means, switched inducers require a power-consuming controller that basic bridges do not. Still, nanowatts for the controller is usually not enough to trump the microwatts that switched inducers gain over basic bridges [21], [30]–[33].

**C. Bridged Switched Inductor**

One way to synchronize discharges into \( \text{v}_{\text{BAT}} \) is to rectify \( \text{v}_{\text{PZ}} \) across half cycles with a bridge and drain \( C_{\text{PZ}} \) into \( \text{v}_{\text{BAT}} \) between half cycles with \( L_X \) like Fig. 14 shows [34]. This way, \( \text{v}_{\text{REC}} \) in Fig. 15 follows \( \text{v}_{\text{PZ}} \) across \( i_{\text{PZ}} \)'s positive half cycles to peak at \( \Delta \text{v}_{\text{PZ}(\text{OC})} \). When half cycles end, switch \( S_G \) closes long enough to drain \( C_{\text{PZ}} \) into \( L_X \). When \( S_G \) opens, \( S_0 \) closes and \( D_G \) conducts to deplete \( L_X \) into \( \text{v}_{\text{BAT}} \). \( \text{v}_{\text{REC}} \) then mirrors \( \text{v}_{\text{PZ}} \) across \( i_{\text{PZ}} \)'s negative half cycles to peak at \( \Delta \text{v}_{\text{PZ}(\text{OC})} \) at which points \( S_G \), \( D_G \), and \( S_0 \) drain \( C_{\text{PZ}} \) into \( L_X \) and \( L_X \) into \( \text{v}_{\text{BAT}} \).

**Fig. 14.** Bridged switched inductor.

To reduce conduction losses, cross-coupled NFETs can replace \( D_{\text{NG}} \) and \( D_{\text{PG}} \) when \( \Delta \text{v}_{\text{PZ}(\text{OC})} \) is greater than an NMOS threshold voltage. Plus, since no component loads \( \text{v}_{\text{REC}} \) across half cycles, cross-coupled PFETs can also replace \( D_{\text{PO}} \) and \( D_{\text{NO}} \) when \( \Delta \text{v}_{\text{PZ}(\text{OC})} \) is greater than a PMOS threshold voltage [35]. If \( \Delta \text{v}_{\text{PZ}(\text{OC})} \) is high enough, \( C_{\text{PZ}} \) can drain into \( L_X \) and \( \text{v}_{\text{BAT}} \) with \( S_{\text{REC}} \) in Fig. 16. Then \( L_X \) can then deplete into \( \text{v}_{\text{BAT}} \) with \( D_G \) [36]. The tradeoff for fewer switches is that \( \Delta \text{v}_{\text{PZ}(\text{OC})} \) must exceed \( \text{v}_{\text{BAT}} \) for the circuit to work.

**D. Bridgeless Switched Inductor**

\( L_X \) in Fig. 17 discharges \( C_{\text{PZ}} \) directly without a bridge [37]. Here, \( i_{\text{PZ}} \) charges \( C_{\text{PZ}} \) across half cycles like Fig. 13 shows. At \( \text{v}_{\text{PZ}} \)'s positive peak \( \Delta \text{v}_{\text{PZ}(\text{OC})} \), \( S_N \) and \( S_p \) close long enough to drain \( C_{\text{PZ}} \) into \( L_X \). Then, \( S_p \) opens and \( L_X \)'s \( \text{i}_L \) flows through \( D_P \).
into \( v_{\text{BAT}} \) until \( L_X \) depletes. At the end of the negative half cycle, when \( v_{pz} \) peaks at \(-\Delta V_{PZ\text{(OC)}}\), \( S_N \) and \( S_P \) again close to drain \( C_{PZ} \) into \( L_X \). Then, \( S_N \) opens and \( L_X \)‘s \( i_x \), which is now flowing up toward \( v_{\text{SW}_-} \), flows through \( D_N \) into \( v_{\text{BAT}} \).

Fig. 17. Bridgeless switched inductor.

Fixing \( C_{PZ} \)‘s bottom terminal to ground creates one subtle, though not insignificant disadvantage. The drawback is, \( C_{PZ} \)‘s negative half-cycle voltage exposes \( S_N \) to negative voltages. This means, conventional CMOS switches must bias their P-type substrates to a voltage that is at least just as negative to isolate the switches from other devices in the die.

The circuit also suffers from one limitation worth noting. For \( D_N \) not to conduct when \( C_{PZ} \) drains into \( L_X \) at the end of \( i_{pz} \)‘s positive half cycle, \( v_{pz} \)‘s peak \( \Delta V_{PZ\text{(OC)}} \) should not exceed \( v_{\text{BAT}} \). This is usually not a problem for tiny transducers because they typically capture a very small fraction of the mechanical energy available.

V. PRE-DAMPING SWITCHED INDUCTORS

A. Pre-Damping

Fortunately, the power \( i_{pz} \) produces climbs with \( v_{pz} \). As a result, pre-charging \( C_{PZ} \) between half cycles and allowing \( i_{pz} \) to charge \( C_{PZ} \) above that level across half cycles, like Fig. 18 illustrates, draws more power from motion than without pre-charging \( C_{PZ} \) [38]. In other words, the system recovers much more than just the energy invested \( E_P \) to pre-damp \( C_{PZ} \) to \( v_P \).

Fig. 18. Synchronized and pre-damped discharges.

To pre-damp \( C_{PZ} \) to \( v_P \) between half cycles, the system must first deposit \( E_P \) or \( 0.5C_{PZ}v_P^2 \) into \( C_{PZ} \). But since \( i_{pz} \) charges \( C_{PZ} \) another \( \Delta V_{PZ\text{(OC)}} \) across each half cycle, the system recovers \( E_{\text{CIPK}} \) or \( 0.5C_{PZ}(v_P + \Delta V_{PZ\text{(OC)}})^2 \). So across an entire cycle, \( v_{\text{BAT}} \) invests and recovers twice these amounts to net \( E_{HI} \):

\[
E_{HI} = 2\left( E_{\text{CIPK}} - E_P \right) = 2\left[ 0.5C_{PZ}(v_P + \Delta V_{PZ\text{(OC)}})^2 - 0.5C_{PZ}v_P^2 \right] = C_{PZ}\Delta V_{PZ\text{(OC)}}^2 + 2C_{PZ}\Delta V_{PZ\text{(OC)}}v_P.
\]

Not surprisingly, \( E_{HI} \) here exceeds that of basic switched inductors. Plus, \( E_{HI} \) grows with \( v_P \), so \( v_P \) should be as high as breakdown voltage \( v_{BD} \) and power losses allow. This means, \( v_P + \Delta V_{PZ\text{(OC)}} \) should not surpass \( v_{BD} \), which is equivalent to saying \( v_P \) should remain below \( v_{BD} - \Delta V_{PZ\text{(OC)}} \).

B. BRIDGED SWITCHED INDUCTORS

\( L_P \) and \( L_P \) in the bridged switched inductor of Fig. 19 pre-damp \( C_{PZ} \) to \( v_P \) between half cycles, like just described, to increase the voltage with which \( i_{pz} \) sources power. [39] does pretty much the same, but with more switches. In all, the bridge rectifies \( v_{pz} \) across half cycles so \( L_X \) can drain \( C_{PZ} \) into \( v_{\text{BAT}} \) and \( L_P+ \) and \( L_P- \) can pre-damp \( C_{PZ} \) between half cycles. So just before a positive half cycle, \( S_c \) closes to energize both \( L_P- \) and \( C_{PZ} \). When \( v_{pz} \) surpasses \( v_{\text{BAT}} \), \( L_P- \) begins to drain into \( C_{PZ} \) what \( L_P+ \) collected before \( v_{pz} \) reached \( v_{\text{BAT}} \). The controller then opens \( S_c \), when \( L_P- \) depletes, at which point \( v_{pz} \) is at \( v_P \).

Fig. 19. Bridged switched inductor.

When the positive half cycle ends, when \( v_{pz} \) peaks to \( v_P + \Delta V_{PZ\text{(OC)}} \), \( S_{REC} \) closes to drain \( C_{PZ} \) into \( L_X \) and \( v_{\text{BAT}} \), and when \( v_{pz} \) falls below \( v_{\text{BAT}} \), \( C_{PZ} \) and \( L_X \) deplete into \( v_{\text{BAT}} \). If \( L_X \) still has energy when \( v_{pz} \) is zero, \( S_{REC} \) opens and \( D_N \) steers what remains in \( L_X \) into \( v_{\text{BAT}} \). Also at this point, \( S_c \) and \( L_P- \) mirror the action of \( S_c \) and \( L_P+ \), to pre-charge \( C_{PZ} \) to \(-v_P \). Therefore closes to energize \( L_P- \) and \( C_{PZ} \) in the negative direction, \( L_P- \) begins to drain into \( C_{PZ} \) when \(-v_{pz} \) surpasses \( v_{\text{BAT}} \), and when \( L_P- \) depletes, \( S_N- \) opens.

Notice \( M_{PG} \) and \( M_{NG} \) steer current into ground when \( S_c \) and \( S_c \) pre-damp \( C_{PZ} \). This means, diodes cannot replace \( M_{PG} \) and \( M_{NG} \). Cross-coupled PFETs, however, can replace \( D_{PO} \) and \( D_{NO} \) because \( S_{REC} \) keeps the FETs from conducting when \( v_{pz} \) engages them (before reaching \( \Delta V_{PZ\text{(OC)}} \)). But since \( D_{PO} \) and \( D_{NO} \) are in series with \( S_{REC} \) switches in place of \( D_{PO} \) and \( D_{NO} \) can incorporate the functionality of \( S_{REC} \). In other words, removing \( S_{REC} \) and replacing \( D_{PO} \) and \( D_{NO} \) with switches that a controller commands is possible. This way, their combined series resistance and losses are lower.

This system suffers from one subtle limitation. For \( L_P- \) and \( L_P+ \) to drain, their voltages must reverse polarity, which can only happen if \( v_P \) surpasses \( v_{\text{BAT}} \). In other words, this circuit works when the optimal pre-damping voltage is higher than \( v_{\text{BAT}} \). This, however, is not a significant limitation when the transducer is under-damped, the breakdown voltage is higher than \( v_{\text{BAT}} \), and incremental losses when \( v_P \) is above \( v_{\text{BAT}} \) do not cancel additional gains.

C. RECYCLING AND BRIDGED SWITCHED INDUCTORS

The switched inductors and bridge in Fig. 20 similarly pre-damp \( C_{PZ} \) after synchronized discharges [40]. Here, the bridge rectifies across half cycles, \( L_X \) drains \( C_{PZ} \) partially into \( v_{\text{BAT}} \), and \( L_R \) recycles what is left in \( C_{PZ} \) back into \( C_{PZ} \) to pre-damp \( C_{PZ} \) in the opposite direction. So when \( v_{pz} \) peaks at \( v_P + \Delta V_{PZ\text{(OC)}} \), \( S_{REC} \), \( D_G \), and \( L_X \) drain \( C_{PZ} \) partially to \( v_P \). \( S_R \), then depletes \( C_{PZ} \) into \( L_R \), and when \( S_R \) still closed, \( L_R \) returns \( C_{PZ} \)’s energy to pre-damp \( C_{PZ} \) to \(-v_P \). At the next half cycle, \( S_{REC} \),
D, and LX similarly drain C PZ from \(-v_P + \Delta v_{PZ(O)}\) to \(-v_P\) and SR and LR recycle what is left to pre-charge C PZ to v P.

To follow the flow of energy into v BAT more closely, the bridge and SREC first discharge C PZ into L X and v BAT until v PZ falls to v P. SREC then opens, so L X’s i L flows from ground to v BAT through D G until L X exhausts its energy into v BAT. The same happens at the end of the negative half cycle: SREC discharges C PZ into L X and v BAT until v PZ is \(-v_P\) and D G then depletes L X into v BAT. In other words, v BAT receives part of the energy C PZ collects across half cycles when SREC closes and the rest when SREC opens. And LR recycles the energy required to pre-damp C PZ to v P and \(-v_P\).

If \(\Delta v_{PZ(O)}\) exceeds the threshold voltages of N- and P-type MOSFETs, cross-coupled NFETs can replace D NG and D P and cross-coupled PFETs can replace D PO and D NO. Since D PO and D NO are in series with SREC, switches in place of D PO and D NO can incorporate the functionality of SREC. Removing SREC and replacing D PO and D NO with switches that a controller commands is therefore possible.

One limitation worth noting is \(v_P + \Delta v_{PZ(O)}\) must exceed v BAT for C PZ to drain into L X and v BAT. In other words, pre-damping must be high enough and vibrations strong enough to charge C PZ to a level that is higher than v BAT. Like before, this is not a problem when the transducer is under-damped, breakdown voltage is higher than v BAT, and the incremental losses of a higher \(v_P\) do not cancel additional gains. Using S G (and SREC) in Fig. 21 [33] to deplete C PZ into L X and D O (with D G) to later drain L X into v BAT circumvents this limitation. The tradeoff for removing this limitation is more switches.

**D. Switched-Inductor Bridge**

Instead of relegating pre-damping and harvesting tasks to separate inductors, Fig. 22 reconciles those two functions into one [41]. So when v PZ peaks, S PZ closes to drain C PZ into L X and v BAT. Then, when v PZ falls below v BAT, both C PZ and L X drain into v BAT. And finally, when v PZ reaches zero, leftover energy in L X pre-damps C PZ to a negative voltage, after which D PO and S PZ open. At the next half cycle, the sequence repeats, but in reverse: C PZ drains into L X and v BAT, both C PZ and L X drain into v BAT, and then leftover energy in L X pre-damps C PZ.

Since L X draws and drains the same energy with mirrored energizing and de-energizing voltages, L X drains with ground and v BAT what L X draws with 2v BAT at C PZ and v BAT. This means, C PZ and L X both exhaust at 0 V when \(\Delta v_{PZ(O)}\) is twice v BAT. Without leftover energy in L X, however, the system discharges, but does not pre-damp C PZ between half cycles to behave like basic switched inductors, like Fig. 13 shows. So to pre-damp C PZ, \(\Delta v_{PZ(O)}\) must exceed 2v BAT.

But with leftover energy used to pre-damp C PZ, v PZ surpasses \(\Delta v_{PZ(O)}\) in the next half cycle to produce even more leftover energy. Leftover energy therefore rises over time to produce the growing oscillations in Fig. 23. v PZ stops swelling when the rise in drawn power in the mechanical domain balances mechanical damping forces, or when the rise in conduction losses cancels the rise in drawn power. In other words, output power climbs until the transducer is no longer under-damped or power losses outpace gains.

**Finding the maximum power point this way, automatically, is appealing when breakdown voltage and other power losses do not limit output power first. Unfortunately, over-damping a tiny transducer is very difficult. So if unchecked, either the circuit exceeds its breakdown limit or incremental power losses in the system outpace additional gains before conduction losses alone do. In other words, either the circuit breaks or outputs less power than possible.**

When \(\Delta v_{PZ(O)}\) is less than 2v BAT, L X depletes before C PZ, so output diodes open and i PZ finishes draining C PZ. i PZ therefore has less charge with which to energize C PZ, so end-of-cycle energy is lower than in the previous cycle. With less initial energy, i PZ loses more charge, so subsequent half cycles output less charge to produce the fading oscillations in Fig. 24.

Since D PO and D NO are in series with S PZ, switches in place...
of \( D_{\text{PO}} \) and \( D_{\text{NO}} \) can incorporate the functionality of \( S_{\text{PO}} \). So removing \( S_{\text{PO}} \) and replacing \( D_{\text{PO}} \) and \( D_{\text{NO}} \) with switches that a controller commands is possible. And if \( \Delta V_{\text{POZO}} \) exceeds an NMOS threshold voltage, cross-coupled NFETs can also replace \( D_{\text{NG}} \) and \( D_{\text{PG}} \).

Incorporating \( S_{\text{PO}} \) into \( D_{\text{PO}} \) and \( D_{\text{NO}} \) in the guise of \( S_{\text{PO}} \) and \( S_{\text{NO}} \) and replacing \( D_{\text{NG}} \) and \( D_{\text{PG}} \) with the synchronous switches in Fig. 25 avails other switching configurations that can ensure lossless oscillations always grow [42]. The network can also limit how much energy drains into \( v_{\text{BAT}} \) when \( \Delta V_{\text{POZO}} \) is less than \( 2v_{\text{BAT}} \) to ensure \( L_x \) has sufficient leftover energy to pre-damp \( C_{\text{PO}} \) to \( v_p \) and \( -v_p \) like Fig. 18 illustrates. For this, \( S_{\text{PO}} \) and \( S_{\text{PG}} \) can, like before, drain \( C_{\text{PO}} \) into \( L_x \) and \( v_{\text{BAT}} \). But unlike [42], \( S_{\text{PO}} \) opens when \( C_{\text{PO}} \) and \( L_x \) hold enough energy to pre-damp \( C_{\text{PO}} \). So when \( S_{\text{NO}} \) closes, \( C_{\text{PO}} \) depletes into \( L_x \) and \( v_{\text{BAT}} \), \( S_{\text{NO}} \) opens and \( S_{\text{PG}} \) closes to pre-damp \( C_{\text{PO}} \) into \( L_x \) and \( v_{\text{BAT}} \).

Interestingly, \( t_{\text{PO}} \) and \( L_x \) limit how much energy drains into \( v_{\text{BAT}} \) when \( \Delta V_{\text{POZO}} \) is less than \( 2v_{\text{BAT}} \) to ensure \( L_x \) has sufficient leftover energy to pre-damp \( C_{\text{PO}} \) to \( v_p \) and \( -v_p \) like Fig. 18 illustrates. For this, \( S_{\text{PO}} \) and \( S_{\text{PG}} \) can, like before, drain \( C_{\text{PO}} \) into \( L_x \) and \( v_{\text{BAT}} \). But unlike [42], \( S_{\text{PO}} \) opens when \( C_{\text{PO}} \) and \( L_x \) hold enough energy to pre-damp \( C_{\text{PO}} \). So when \( S_{\text{NO}} \) closes, \( C_{\text{PO}} \) depletes into \( L_x \) and \( v_{\text{BAT}} \) until \( S_{\text{NO}} \) opens and \( S_{\text{PG}} \) closes to pre-damp \( C_{\text{PO}} \) into \( L_x \) and \( v_{\text{BAT}} \).

$$\Delta V_{\text{POZO}} = 2V$$
$$P_0 = 12.0 \mu W$$

In operating, \( \text{cp} \) charges \( C_{\text{PO}} \) across \( \text{cp} \)’s positive half cycles to \( \Delta V_{\text{POZO}} \). At that point, \( v_{\text{BAT}} \) closes to draw pre-damping energy from \( v_{\text{BAT}} \) into \( L_x \). \( S_{\text{PO}} \) then opens and \( S_{\text{PG}} \) closes to deplete \( C_{\text{PO}} \) to \( L_x \). \( S_{\text{PG}} \) remains closed long enough to cycle \( L_x \)’s combined energy back to \( C_{\text{PO}} \). This way, \( C_{\text{PO}} \) pre-damps to a level \(-2v_p \) in Fig. 27 that \( v_{\text{BAT}} \)’s investment controls. \( \text{cp} \) then charges \( C_{\text{PO}} \) across \( \text{cp} \)’s negative half cycle to \(-2(2v_p + \Delta V_{\text{POZO}}) \). At that point, \( S_{\text{PO}} \) first closes to deplete \( C_{\text{PO}} \) into \( L_x \). Once drained, when \( v_{\text{PO}} \) is zero, \( S_{\text{PO}} \) opens and \( S_{\text{BAT}} \) closes to drain \( L_x \) into \( v_{\text{BAT}} \). This sequence then repeats.

As with its bridgeless predecessor from Fig. 17, \( C_{\text{PO}} \)’s negative half-cycle voltage exposes \( S_{\text{PO}} \) to negative voltages. Conventional CMOS switches must therefore bias their P-type substrates to a voltage that is at least just as negative to isolate the switches from other devices in the die. This is a subtle, though not insignificant requirement for this circuit.

**F. Recycling Switched-Inductor Bridge**

The switched-inductor in Fig. 28 adopts a different approach [44]. Here, \( L_R \) recycles the charge that \( \text{cp} \) loses the first time \( \text{cp} \) charges \( C_{\text{PO}} \) across the bridge’s \( 2V_{\text{REC}} \) so \( V_{\text{REC}} \) can receive all of \( \text{cp} \)’s charge after that. For this, \( S_R \) closes between half cycles to drain \( C_{\text{PO}} \) into \( L_R \) and \( L_R \) back into \( C_{\text{PO}} \) and swing \( v_{\text{PO}} \) in Fig. 29 from \( v_{\text{REC}} \) to \(-v_{\text{REC}} \) at the end of the positive half cycle and back from \(-v_{\text{REC}} \) to \( v_{\text{REC}} \) at the end of the other half. Since \( v_{\text{PO}} \) is already at \( v_{\text{REC}} \) before new half cycles begin, all of \( \text{cp} \)’s charge flows into \( V_{\text{REC}} \).

This strategy features two important traits. First, \( V_{\text{REC}} \) collects all \( \text{cp} \)’s charge. Second, since \( \text{cp} \) no longer loses charge to \( C_{\text{PO}} \), \( v_{\text{PO}} \)’s loaded swing \( \Delta V_{\text{POZO}} \) need no longer halve \( v_{\text{PO}} \)’s unloaded counterpart \( \Delta V_{\text{POZO}} \). As a result, no tradeoff counters the rise in power that \( \text{cp} \) produces at \( v_{\text{REC}} \) when \( V_{\text{REC}} \) is higher. \( V_{\text{REC}} \) can therefore collect twice \( \text{cp} \)’s half-cycle charge \( q_{\text{HALF}} \) at the highest possible \( V_{\text{REC}} \) all the time:

$$E_{\text{H}} = 2q_{\text{HALF}} v_{\text{REC}} = 2C_{\text{PO}}(\Delta V_{\text{POZO}}) v_{\text{REC}} \leq 2C_{\text{PO}}(\Delta V_{\text{POZO}}) V_{\text{BD}}.$$  \((15)\)

But since \( v_{\text{BAT}} \) is seldom at the breakdown voltage \( V_{\text{BD}} \), the recycling diode bridge cannot charge \( v_{\text{BAT}} \) directly. So like its predecessor, the circuit requires a maximum power-point charger. The basic aim of the charger is, like before, to draw just enough power to keep \( v_{\text{REC}} \) near \( V_{\text{BD}} \).

**VI. COMPARISON**

**A. Output Power**

Actual output power is often more the result of application and implementation than fundamental principles. Space and process technologies, for example, dictate inductor’s resistance, switches’ resistances and capacitances, and controller’s operating power, almost none of which are
uniform in literature. This is why comparing output power or conversion efficiencies of reported implementations can be misleading when starting a design. Comparing possibilities and guiding principles are more important in this respect.

Half bridges generate as much power as full bridges because, at their maximum power points, they collect half the charge at twice the voltage. Basic switched inductors, however, collect all the charge to output four times the energy that basic bridges can. But since under-damped piezoelectric transducers source more power with higher voltages, pre-damping switched inductors, which pre-charge \( C_{\text{pZ}} \) between half cycles, output even more power. In fact, the recycling switched-inductor bridge produces the highest power because \( C_{\text{REC}} \) collects \( i_{\text{pZ}} \)'s charge at the highest possible voltage all the time.

Pre-damping circuits, however, can only output as much power as their breakdown voltages allow. Still, under similar loaded-swing limits, the recycling bridge outputs more power. With \( \pm 9.4 \) \( \mu \)A, 15 nF, and a 6-V limit, for example, all other pre-damping switched inductors output 40% less power at 12 \( \mu \)W than the recycling bridge at 20 \( \mu \)W, as the lossless simulations in Figs. 17, 26, and 28 demonstrate.

Power losses, however, offset these gains to, in some cases, negate their benefits. Although controllers can nowadays consume microwatts or less [21], [30]-[33], switches in the power-conduction path can lose tens of microwatts or more to gate drive and ohmic power [43]. This is why bridgeless schemes can, with only two to four switches, out-power their in-class competitors. The recycling bridge, for example, like all other bridges, requires a charger that draws just enough power to keep its rectified output at its maximum power point. But since chargers typically rely on at least two switches to transfer power [46], seven switches can burn the additional power that the recycling bridge can produce. Plus, the synchronizing controller that switches the inductor and the buffer that regulates the intermediate rectified voltage require more quiescent power than either on its own, which means quiescent power for the recycling bridge is also higher.

These conclusions hinge on under-damped conditions to persist. If harvesting chargers over-damp transducers before they reach their breakdown limits, the power-producing benefits of pre-damping systems disappear. What matters then are power losses, which means the bridgeless two-switch configuration will probably output the highest power.

The challenge with the two-switch topology is \( V_{\text{BAT}} \) only collects energy after negative half cycles. This means, the system damps the transducer across negative half cycles more than across positive half cycles, so the cantilever bends less in that direction than in the other. This may not be a problem when heavily under-damped because damping effects on displacement can be negligible. But when nearing overdamped conditions, when the two-switch solution wins over competing technologies, asymmetrical damping may compromise the mechanical stability of the transducer.

Near over-damped conditions, the next least power-consuming system is the four-switch bridgeless sibling. This one, however, like its two-switch counterpart, requires a negative supply, which again, consumes power. But since the purpose of the negative supply is to bias the substrate, it does not transfer much power, so power losses can be low.

**B. Integration**

When considering microsystems, reducing the number of off-chip components is paramount. Although basic bridges do not require regulating chargers to output power, they cannot operate at their maximum power point without the chargers. Efficient chargers, however, use at least one off-chip inductor to transfer power [46]. So all harvesters in the state of the art ultimately require no less than one inductor, as Table II shows.

But since basic bridges and the recycling bridge also require a rectifying capacitor, basic switched inductors, the switched-inductor bridge, and the bridgeless pre-damping switched inductor require less board space (for just one inductor). Of these, with only two switches, the pre-damping bridgeless scheme occupies the least silicon area. For reference, [38] used a 2.7-cm cantilever that integrates 25.4 \( \times \) 3.8 \( \times \) 0.25 mm\(^3\) of piezoelectric material and a 330-\( \mu \)H surface-mount inductor with 1.6 \( \Omega \) of series resistance that occupies 6 \( \times \) 6 \( \times \) 3.5 mm\(^3\).

**VII. Conclusions**

Of prevailing and derived technologies, the recycling bridge outputs the highest power because the transducer outputs charge that the system collects at the highest possible voltage all the time. The pre-damping bridgeless switched inductor is next, and with only one inductor and two switches, this bridgeless option occupies less board space and less silicon.

<table>
<thead>
<tr>
<th>Table II. Comparison of the State of the Art</th>
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<tr>
<td><strong>Basic Bridges</strong></td>
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<tr>
<td><strong>Full</strong></td>
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<td>Collected Charge</td>
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<td>Max. Energy/Cycle</td>
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<td>Inductors</td>
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<td>( C_{\text{REC}} )</td>
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<tr>
<td>Negative Supply</td>
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<td>Limitations</td>
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Buffer normally requires one inductor and additional switches. \(^1\)Removing this limitation requires one switch. \(^2\)Removing this limitation requires two switches.
area. Plus, when over-damped conditions limit draw power, this two-switch solution consumes less power, so it outputs more power than the recycling diode bridge. But if asymmetrical damping compromises the mechanical stability of the transducer, the four- and five-switch bridgeless sibling and synchronous switched-inductor bridge are better options. Since under-damped conditions prevail in most micro-scale applications, however, the benefits of the recycling diode bridge and pre-damping bridgeless options are ultimately difficult to discount.

REFERENCES


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