Harvesting Ambient Kinetic Energy with Switched-Inductor Converters

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Abstract—The potential application space for miniaturized systems like wireless microsensors is expansive, from reconnaissance mission work and remote sensors to biomedical implants and disposable consumer products. Conforming to microscale dimensions, however, constrains energy and power to such an extent that sustaining critical power-hungry functions like wireless communication is problematical. Harvesting ambient kinetic energy offers an appealing alternative, except the act of transferring energy requires power that could easily exceed what the harvester generates in the first place. This paper reviews piezoelectric and electrostatic harvester circuits, describes how to design low-power switched-inductor converters capable of producing net energy gains when supplied from piezoelectric and electrostatic harvesters, and presents experimental results from prototype embodiments. In the electrostatic case shown, the controller dissipated 0.91 nJ per cycle and the switched-inductor precharger achieved 90.3% efficiency to allow the harvester to net a gain of 2.47 nJ per cycle from a capacitor that oscillated between 157 and 991 pF. The piezoelectric counterpart harnessed 1.6 to 29.6 μJ from weak periodic vibrations with 0.05 – 0.16-m/s² accelerations and 65.3 μJ from (impact-produced) non-periodic motion.

Index Terms—Harvesting/harnessing energy, vibration, piezoelectric, electrostatic, wireless microsensor, battery charger, rectifier, switched-inductor converter.

I. HARVESTING KINETIC ENERGY IN VIBRATIONS

Wireless microsensors can enjoy popularity in, for example, biomedical implants [1] and tire-pressure monitoring systems [2] because they offer in-situ, real-time, non-intrusive processing capabilities, except miniaturized platforms limit the energy onboard batteries can store, so lifetimes are short. Ambient energy is an attractive alternative because harnessing energy from light, heat, RF radiation, and motion can continuously replenish an exhaustible reservoir.

Of these sources, solar light produces the highest power density, except when supplied from indoor lighting under which conditions power decreases drastically [3]. Harnessing thermal energy is viable, but microscale dimensions severely limit temperature gradients, the fundamental mechanism from which thermopiles draw power [4]–[5]. Mobile electronic devices today radiate plenty of RF energy, but still, power drops with distance to impracticable levels [6]. Harvesting kinetic energy may not compete with solar power, but in contrast to indoor lighting, thermal, and RF sources, moderate and consistent vibration power across a vast range of applications is typical [7]–[9].

Although operating conditions ultimately determine which kinetic energy-harvesting method is optimal, piezoelectric transducers are relatively mature and produce comparatively higher power than their counterparts [9]. On-chip piezoelectric devices, however, remain the subject of ongoing research [10]–[12], which is where electrostatic harvesters find an edge, because MEMS technologies can more aptly integrate variable, parallel-plate capacitors on chip [8].

Piezoelectric and electrostatic transducers harvest ambient kinetic energy by converting mechanical energy in vibrations (E_{ME} in Fig. 1) into the electrical domain (E_{EE}). More specifically, piezoelectric bimorph strips produce charge when bent and parallel-plate capacitors when their plates separate. To harness and store the generated E_{EE}, harvester circuits condition transducers and deliver a net energy gain E_{NET} to intermediate reservoirs that can supply power to electrical loads on demand. Harvesters do not supply energy to the load directly because the mechanical input is often unpredictable and therefore unreliable for sporadic loading events [13].

![Diagram of harvesting system](image)

Fig. 1. Harvesting ambient kinetic energy in vibrations.

To generate E_{NET} from miniscule amounts of E_{EE}, harvesters must dissipate little. For this reason, switched-inductor circuits, which are already popular in low-power microelectronics because of their high efficiency levels, match the requirements of the harvesting function well. In addition, as this paper shows, switched-inductor converters can also condition and induce transducers to generate more E_{EE} from E_{ME} than their inductor-free and diode-based counterparts.
Accordingly, while [14] briefly introduced and described how switched-inductor circuits can harvest kinetic energy, this paper details the operation and prototype implementations of ICs designed to harvest ambient kinetic energy in vibrations with piezoelectric and electrostatic transducers. To that end, Sections II and III establish the basic harvesting theory on which illustrative piezoelectric and electrostatic switched-inductor converter examples presented rely to generate a net energy gain. Section IV then explains how these circuits dissipate (lose) power and Section V evaluates how the prototyped systems performed experimentally. The paper ends by drawing relevant conclusions in Section VI.

II. PIEZOELECTRIC HARVESTERS

A. Piezoelectric Transducers

When a mechanical vibration bends a piezoelectric material, the stress rearranges the internal lattice structure to shift the charge balance of the crystal [15]. Fig. 2a, for example, illustrates piezoelectricity in an ionic non-centrosymmetric crystal like quartz [16]. Here, stress shifts the centers of the positive (cations) and negative (anions) charges in opposite directions to produce a surface potential. The potential and its associated currents change continuously with variations in mechanical deformation. In essence, the material behaves as an ac current source (iPZT in Fig. 2b) that charges and discharges the capacitance (C\text{PZT}) across the surfaces of the material, where a leakage (R\text{LEAK}) represents a slight drain [9].

While typical piezoelectric transducers, which are relatively mature and reliable technologies, can produce high peak voltages [17], their miniaturized counterparts cannot [10]–[12], because the amount of charge generated is proportional to strain and surface area, both of which decrease considerably with reductions in volume. Furthermore, typical environments supply mostly "weak" vibrations, energy (and voltage) from which is so low that harvesters can hardly operate.

B. Energy-Harvesting Systems

Rectifier Based: To rectify and channel the ac power that a piezoelectric transducer generates into an intermediate dc storage device, like a battery or capacitor, harvesters often employ full-wave diode-bridge rectifiers [18]–[20]. These rectifier circuits, as Fig. 3 exemplifies, steer charge to the output only when iPZT charges C\text{PZT} above the barrier voltage that two conducting diodes (2V\text{D}) and an output capacitor C\text{RECT} (V\text{RECT}) produce. Since harvesting more energy amounts to channeling more of iPZT into C\text{RECT}, C\text{RECT} is, by design, substantially larger than C\text{PZT}. Accordingly, in one vibration cycle T\text{VIB}, V\text{RECT} remains practically unchanged and iPZT clamps to the aggregate sum of 2V\text{D} and V\text{RECT} when iPZT flows into C\text{RECT} across conduction time T\text{COND}.

Because leakage in piezoelectric transducers is typically negligible (i.e., R\text{LEAK} is large) and C\text{PZT} is a reactive component (which does not consume power), most of the energy iPZT carries through the diodes reaches the output, except for the power the diodes dissipate. Rectifier efficiency \( \eta_{\text{RECT}} \), the ratio of rectified output energy per cycle E\text{RECT} to the input energy of the rectifier per cycle E\text{EE}, therefore deteriorates with increasing diode voltage V\text{D}:

\[
\eta_{\text{RECT}} = \frac{E_{\text{RECT}}}{E_{\text{EE}}} = \frac{2 \int_{i_{\text{PZT}}}^{i_{\text{PZT}} = 0} (V_{\text{RECT}} + 2V_{\text{D}}) \, dt}{\int_{i_{\text{PZT}}}^{i_{\text{PZT}} = 0} (V_{\text{RECT}} + 2V_{\text{D}}) \, dt} = \frac{V_{\text{RECT}}^2 + 2V_{\text{D}}}{V_{\text{RECT}}^2 + 2V_{\text{D}}}.
\]

Replacing the diodes with synchronous MOS switches increases \( \eta_{\text{RECT}} \) because MOSFETs reduce V\text{D} from 0.5 – 0.7 V to mVs [19]–[22]. Still, a fundamental limitation with rectifiers is that vPZT must exceed V\text{RECT} (which represents the energy already stored in C\text{RECT}) to extract energy from motion. In other words, even if V\text{D} were zero, the rectifier stops harvesting when vPZT’s open-circuit peak voltage V\text{P} falls below V\text{RECT}, which can easily happen under weak vibrations.

One way of reducing the input threshold of the rectifiers is by conditioning C\text{RECT} so V\text{RECT} can remain below vPZT’s peak voltage [18]. This way, the system can harness some of the energy C\text{PZT} receives under weak vibrations that the transducer would have otherwise transferred back into the mechanical domain. To quantify the E\text{RECT} as a function of V\text{RECT}, consider iPZT would charge C\text{PZT} from −V\text{P} to V\text{P} with \( \Delta Q_{\text{PZT}} \) without the rectifier. When the rectifier loads the transducer, because some portion of \( \Delta Q_{\text{PZT}} \) is used only to charge up C\text{RECT} from −(V\text{RECT} + 2V\text{D}) to (V\text{RECT} + 2V\text{D}) as shown in Fig. 3, the rectifier steers \( \Delta Q_{\text{RECT}} \):

\[
\Delta Q_{\text{RECT}} = \Delta Q_{\text{PZT}} + C_{\text{PZT}}[2(V_{\text{RECT}} + 2V_{\text{D}})].
\]

(2)

into C\text{RECT} (at V\text{RECT}) for each half-cycle. Therefore E\text{RECT} grows with V\text{P} and become a second-order function of V\text{RECT}:

\[
E_{\text{RECT}} = 2(V_{\text{RECT}}^2 + \Delta Q_{\text{RECT}}) = 4V_{\text{REC}}^2C_{\text{PZT}}[V_{\text{P}} - (V_{\text{RECT}} + 2V_{\text{D}})].
\]

(3)

In other words, the maximum possible energy per cycle a rectifier can harness is \( C_{\text{PZT}}(V_{\text{P}} - 2V_{\text{D}})^2 \), which results when V\text{RECT} is 0.5(V\text{P} – 2V\text{D}). Although monitoring vPZT and adjusting V\text{RECT} accordingly improves performance [18], [23], sensing vPZT and conditioning V\text{RECT} "on the fly" can easily consume more energy than the system can harness under weak vibrations, even if V\text{D} were zero. Here, the complexity of the system also imposes a lower bound on vPZT’s V\text{P} below which the rectifier cannot produce a net energy gain.
Switched Inductor: To circumvent the fundamental input threshold limitation rectifiers impose, the system in Fig. 4 discharges \( C_{\text{PZT}} \) into inductor \( L_{i} \) because \( C_{\text{PZT}} \)’s voltage \( V_{\text{PZT}} \) energizes \( L_{i} \), irrespective of how small \( V_{i} \) is. The switched-inductor circuit finishes the harvesting process by de-energizing \( L_{i} \) into battery \( V_{\text{BAT}} \), which again, poses no threshold limit. As such, the rectifier-free, switched-inductor circuit shown can harness kinetic energy from weak vibrations.

![Fig. 4. Rectifier-free, switched-inductor piezoelectric energy-harvesting cycle.](image)

The harvester in Fig. 4 first waits for vibrations to move the piezoelectric cantilever upward in the positive half-cycle, allowing \( i_{\text{PZT}} \) to charge \( C_{\text{PZT}} \) until \( V_{\text{PZT}} \) peaks. At the peak, the circuit quickly discharges \( C_{\text{PZT}} \) into \( L_{i} \) and then de-energizes \( L_{i} \) into \( V_{\text{BAT}} \). This inverting transfer, where the harvester charges \( V_{\text{PZT}} \) negative \( V_{\text{BAT}} \), concludes the vibration cycle, allowing subsequent cycles to repeat the sequence as long as vibrations persist.

In the positive half cycle, \( C_{\text{PZT}} \) stores the energy produced by the transducer \( (E_{\text{EE}}) \) so, when the switched-inductor converter extracts it, \( C_{\text{PZT}} \) resets to 0 V. Assuming \( i_{\text{PZT}} \) is sinusoidal at \( I_{s} \sin(\omega t) \) and all of it flows into \( C_{\text{PZT}} \), \( V_{\text{PZT}} \) is

\[
V_{\text{PZT}}(t) = \frac{1}{C_{\text{PZT}}} \int i_{\text{PZT}}(t) \, dt = \left[ \frac{I_{s}}{\omega_{\text{EE}} C_{\text{PZT}}} \left( 1 - \cos(\omega_{\text{EE}} t) \right) \right].
\]

Therefore, after the positive half-cycle, \( C_{\text{PZT}} \) stores \( E_{\text{EE}}^+ \):

\[
E_{\text{EE}}^+ = \frac{1}{2} \int_{0}^{T/2} i_{\text{PZT}}(t) V_{\text{PZT}}(t) \, dt = \frac{2 I_{s}^2}{\omega_{\text{EE}} C_{\text{PZT}}} = 2 \frac{C_{\text{PZT}}}{V_{P}^2},
\]

and the system harvests a net energy gain \( E_{\text{NET}}^+ \) that is

\[
E_{\text{NET}}^+ = E_{\text{EE}}^+ - E_{\text{LOSS}}^+,
\]

where \( E_{\text{LOSS}}^+ \) includes conduction, switching, and control-circuit losses across the circuit during the positive half-cycle.

Since the negative cycle generates an equivalent amount, the system theoretically harnesses \( 4 C_{\text{PZT}} V_{P}^2 \) from each period, which is four times more energy than ideal rectifiers can (at \( C_{\text{PZT}} V_{P}^2 \)) under optimal operating conditions when \( V_{D} = 0 \) and \( V_{\text{RECT}} \) is 0.5 \( V_{P} \). Fundamentally, this difference arises because the switched-inductor harvester exhausts \( C_{\text{PZT}} \) to 0 V every half cycle so \( C_{\text{PZT}} \)’s effective peak voltage increases to 2 \( V_{P} \), rather than \( V_{P} \) which means the electrical damping force is higher. In other words, the harvester forces the transducer to generate more \( E_{\text{EE}} \) (i.e., 0.5 \( C_{\text{PZT}} V_{P}^2 \)) from the given \( E_{\text{ME}} \).

Fig. 5 depicts the simplified switched-inductor power stage just described and the corresponding simulation waveforms for \( V_{\text{PZT}} \) and inductor current \( i_{L} \). First, through the positive cycle, switch \( S_{L} \) decouples the power-stage from the transducer until \( V_{\text{PZT}} \) reaches its positive peak. The system then discharges \( C_{\text{PZT}} \) into \( L_{i} \) by engaging switches \( S_{L} \) and \( S_{N} \), until \( i_{L} \) peaks. Since LC resonance drives this energy transfer, the system estimates \( L_{i} \)’s energizing time by waiting for one quarter of \( L_{i}C_{\text{PZT}} \)’s resonance period at 0.5 \( V_{P} \), instead of sensing \( i_{L} \)’s peak directly, which would otherwise require considerable power. After this, \( S_{N} \) opens and \( i_{L} \) charges the parasitic capacitance at switching node \( V_{SW}^+ \) quickly until non-inverting diode-switch \( D_{N} \) forward biases and depletes \( L_{i} \) into \( V_{\text{BAT}} \). Note that, without a harvesting circuit, the unloaded transducer charges \( C_{\text{PZT}} \) to lower voltages \( V_{\text{PZT \,(UNLOADED)}} \) which demonstrates the conditioning effect of the harvester.

![Fig. 5. (a) Switched-inductor power stage and (b) simulated waveforms.](image)

During the negative cycle, \( S_{N} \) decouples the circuit from the transducer until \( V_{\text{PZT}} \) reaches its negative peak. \( S_{L} \) and \( S_{N} \) then discharge \( C_{\text{PZT}} \) into \( L_{i} \) for one quarter of \( L_{i}C_{\text{PZT}} \)’s resonance period. Afterwards, \( S_{L} \) opens and \( D_{L} \) conducts \( i_{L} \) into \( V_{\text{BAT}} \).

In the end, unlike rectifier-based harvesters, the switched-inductor circuit can harness energy from small \( V_{\text{PZT}} \) voltages because \( L_{i} \) energizes and stores energy as soon as \( |V_{\text{PZT}}| \) rises above zero. Additionally, \( L_{i} \), like a current source, automatically raises the switching nodes to whatever voltage \( V_{\text{BAT}} \) defines, removing the extra de-de converter stage that typically bridges the voltage gap between the rectifier and \( V_{\text{BAT}} \). In fact, dc converters and other control circuits in [18] not only buffer the rectifier from the battery but also include a lossy feedback loop to optimally condition \( V_{\text{RECT}} \). In contrast, a switched-inductor harvester can derive four times (4 \( \times \)) more energy than the best diode-bridge rectifier with a relatively simpler (i.e., low-power) control scheme.

C. Synchronization and Control

Fig. 6 illustrates a complete rectifier-free, switched-inductor piezoelectric harvester system. Except for the piezoelectric cantilever, \( L_{i} \), and a battery, integrating all power switches and controller circuits on chip is possible. Switches \( S_{L} \) and \( S_{N} \) consist of two series NMOS transistors in an isolated p well to eliminate the undesirable conduction path the body diode would otherwise establish when using only one NMOS or PMOS switch, because the input \( (V_{\text{PZT}}) \) can swing above \( V_{\text{BAT}} \) and below ground. In the case presented, back-to-back diodes block reverse current because the p-well potential tracks the lower of the two terminal voltages present across the off-state switch. Two series PFETs would also block reverse current, but fully engaging PFETs when \( V_{\text{PZT}} \) is below ground requires
a gate-drive voltage that is well below \(v_{\text{PZT}}\)'s negative peak. In contrast, \(V_{\text{BAT}}\) is sufficiently high to engage NFETs when \(v_{\text{PZT}}\) falls below ground and \(v_{\text{PZT}}\)'s negative peak is sufficiently low to disengage them when \(v_{\text{PZT}}\) rises. Note one of the diodes always charges the \(p\) well to the most negative potential when the switch is off and the FETs short the well to the switch's terminal voltages otherwise, so the circuit always biases the well. With respect to \(D_N\) and \(D_h\), the body terminals of \(M_{\text{PDN}}\) and \(M_{\text{PDI}}\) connect to \(V_{\text{BAT}}\) so their body diodes (i) block reverse (battery-discharge) current and (ii) clamp switching nodes \(v_{SW^+}\) and \(v_{SW^-}\) to a diode voltage above \(V_{\text{BAT}}\).

**Peak-\(v_{\text{PZT}}\) Detector and Peak-\(i_i\) Estimator:** \(R_{PD}C_{PD}C_{PK}\) in Fig. 6 implements a peak-voltage detector that prompts the system to drain \(C_{\text{PZT}}\) into \(L_H\) when vibrations maximally charge \(C_{\text{PZT}}\). Here, comparator \(C_{PK}\) detects when \(v_{\text{PZT}}\) peaks by comparing \(v_{\text{PZT}}\) to its delayed counterpart \(v_D\). To extend \(C_{PK}\)'s input common-mode range below ground (to sense \(v_{\text{PZT}}\)'s negative peak), Schottky diode \(D_S\) and \(C_{SS}\) set negative supply voltage \(V_{SS}\). Note that reducing power in \(C_{PK}\) is crucial because it must detect peaks through the vibration period. Fortunately, the cycle is long, which means designers can trade speed for power by, for example, operating in sub-threshold. With a delay of \(0.5\pi\sqrt{L_H/C_{\text{PZT}}}\), \(t_{\text{DLY}}\) ends the \(C_{\text{PZT}}\)-to-\(L_H\) energy-transfer period. This way, with a priori knowledge of \(L_H\) and \(C_{\text{PZT}}\), the system estimates (rather than sense) \(i_i\)'s peak, thereby trading accuracy (and complexity) in favor of power.

**Self-Synchronized Switch:** As mentioned, because \(D_N\) and \(D_h\) in Fig. 5 rectify inductor currents, there is no \(v_{\text{PZT}}\) threshold below which the system cannot harness energy even if p-n junction diodes implement \(D_N\) and \(D_h\). Still, considerable loss incurs due to their voltage drop. Replacing \(D_N\) and \(D_h\) with \(M_{\text{PDN}}\) and \(M_{\text{PDI}}\) in Fig. 6 can reduce this loss significantly.

Synchronizing \(M_{\text{PDN}}\) and \(M_{\text{PDI}}\), however, requires precise (well-timed) control signals. If \(S_N\) and \(S_h\)'s gate-control signals overlap, for example, with those of \(M_{\text{PDN}}\) and \(M_{\text{PDI}}\), \(S_N\)-\(M_{\text{PDN}}\) and \(S_h\)-\(M_{\text{PDI}}\) might initially short-circuit \(V_{\text{BAT}}\) to ground, or to the input, which is why dead time between adjacent switches is imperative. Unfortunately, a fixed dead time in light of variable mechanical input strength, that is, an unpredictable peak inductor current, also causes additional losses [24]-[25]. With a short dead time, \(M_{\text{PDN}}\) and \(M_{\text{PDI}}\) conduct before \(v_{SW^+}\) and \(v_{SW^-}\) reach \(V_{\text{BAT}}\), which causes the switches to dissipate power. A long dead time, on the other hand, allows \(i_i\) to charge \(v_{SW^+}\) and \(v_{SW^-}\) above \(V_{\text{BAT}}\), increasing the time body diodes conduct and dissipate power. Plus, after charging the battery, disengaging \(M_{\text{PDN}}\) and \(M_{\text{PDI}}\) after \(v_{SW^+}\) and \(v_{SW^-}\) fall below \(V_{\text{BAT}}\), discharges the battery with reverse current.

Fundamentally, the above-mentioned losses result because \(M_{\text{PDN}}\) and \(M_{\text{PDI}}\) transition with non-zero voltages. The system in Fig. 6 reduces this loss by using comparators \(C_{DI}\) and \(C_{PDN}\) to sense when \(v_{SW^+}\) and \(v_{SW^-}\) reach \(V_{\text{BAT}}\), thereby engaging \(M_{\text{PDN}}\) and \(M_{\text{PD}}\) only when their terminals are close to 0 V. In practice, however, \(C_{DI}\) and \(C_{PDN}\) require a finite time to respond, which on one end extends the time the lossy body-diodes conduct charging currents and on the other prompts the switches to drain the battery with reverse current.

A fast response often implies considerable quiescent current. The comparator implementation in Fig. 7 minimizes this overhead by ensuring the circuit operates only when needed: when \(L_H\) has sufficient energy to deliver to \(V_{\text{BAT}}\) and power the comparator. More explicitly, as \(L_H\) de-energizes into \(C_{\text{PAR}}\) to raise \(v_{SW}\), some of \(i_i\) flows into mirror \(M_{\text{NB}}\)-\(M_{\text{NO}}\) to activate the comparator, which then compares the currents that \(M_{\text{PB}}\) and \(M_{\text{PO}}\)'s source-gate voltages generate. Since \(M_{\text{PB}}\) and \(M_{\text{PO}}\) match, \(M_{\text{PB}}\) drives more current and engages \(M_{\text{P}}\) when \(v_{SW}\) is above \(V_{\text{BAT}}\), and vice versa. Since \(L_H\) de-energizes only twice per cycle and the required time is a small fraction of the cycle, comparator losses are low.

Because \(v_{SW}\) is initially zero, \(M_{\text{PD}}\) engages to raise \(v_{CP,\text{OUT}}\) to \(V_{\text{BAT}}\), which ensures \(M_{\text{P}}\) is off. Then, \(L_H\) drives current both through \(M_{\text{PB}}\) and into \(C_{\text{PAR}}\) to power the comparator and raise \(v_{SW}\) quickly, the overdrive of which accelerates the circuit's response time. Once \(v_{SW}\) rises above \(V_{\text{BAT}}\) and \(M_{\text{PB}}\)-\(M_{\text{NB}}\)-\(M_{\text{NO}}\) lowers \(v_{CP,\text{OUT}}\), \(M_{\text{P}}\) engages and \(v_{SW}\) remains above \(V_{\text{BAT}}\) by the Ohmic voltage \(i_i\) produces across \(M_{\text{P}}\)'s series resistance.

As a result, \(v_{SW}\) decreases with \(i_i\) as \(L_H\) de-energizes into \(V_{\text{BAT}}\). When \(i_i\) reaches zero, which happens when the system depletes \(L_H\)'s energy, \(v_{SW}\) drops to \(V_{\text{BAT}}\) and the comparator transitions to disengage \(M_{\text{P}}\). Because \(i_i\) no longer carries sufficient current to power the comparator, the circuit shuts off automatically (i.e., synchronously).

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**Fig. 6.** Switched-inductor piezoelectric energy-harvesting system.

**Fig. 7.** Self-synchronized MOS switch \(M_s\) emulates an asynchronous diode.

During the turn-off process, the overdrive input voltage is small (because \(v_{SW}\) is near \(V_{\text{BAT}}\)) so the shut-off operation is relatively slow, which means \(M_{\text{P}}\) can conduct reverse current momentarily. To prevent this, the comparator raises its shut-off trip-point by leaking current away from \(M_{\text{NB}}\) through \(M_{\text{SNB}}\).
so that the circuit switches earlier, when $v_{SW}$ is slightly above $V_{BAT}$. In addition to this offset for shut-off transition, $M_{SH2}$’s current establishes hysteresis for the turn-on transition that helps keep transient and ringing events in $v_{SW}$ from asserting inadvertent transitions in $v_{CP\_OUT}$.

III. ELECTROSTATIC HARVESTERS

A. Electrostatic Transducers

An electrostatic harvester harnesses energy from the work vibrations exert against the electrostatic force of a motion-sensitive, parallel-plate variable capacitor $C_{VAR}$ [7]–[8]. As motion separates $C_{VAR}$’s plates, capacitance decreases and either $C_{VAR}$’s voltage $v_C$ rises to increase its stored energy $E_C$ or charge $q_C$ decreases to generate current $i_{HARV}$ as $\Delta q_C/\Delta t$.

Fabricated variable capacitors from microelectromechanical systems (MEMS) technologies in literature typically feature capacitances that range between 50 and 400 pF [26]–[30] with a maximum $\Delta C_{VAR}$ of 62 to 1570 pF [31]. In practice, these capacitors exhibit parasitic series resistances (across each parallel plate) that dissipate power and capacitances (between plates and substrate/sidewalls) that require additional charge (i.e., energy) from the system.

B. Energy-Harvesting Systems

Charge- vs Voltage-Constrained: When constraining $C_{VAR}$’s charge $q_C$ (by leaving the charged $C_{VAR}$ open-circuited) while vibrations separate $C_{VAR}$’s parallel plates, capacitance decreases and capacitor voltage $v_C$ rises, augmenting the energy stored in $C_{VAR}$ with kinetic energy from ambient vibrations. Since charge remains constant, the capacitance variation effectively amplifies $C_{VAR}$’s initial voltage $V_{C(INI)}$ by maximum-minimum ratio $C_{MAX}/C_{MIN}$ to final value $V_{C(FIN)}$:

$$V_{C(FIN)} = \frac{C_{MAX}}{C_{MIN}} V_{C(INI)}.$$  \hspace{1cm} (7)

Therefore, the net energy gain stored in $C_{VAR}$ (which refers to converted mechanical energy) across a single cycle (i.e., one $\Delta C_{VAR}$ transition) is the difference between final and initial energy stored in the capacitor $E_{C(FIN)}$ and $E_{C(INI)}$:

$$E_{GAIN} = E_{C(FIN)} - E_{C(INI)} = \frac{0.5 C_{MAX}}{C_{MIN}} \Delta C_{VAR} V_{C(INI)}^2.$$  \hspace{1cm} (8)

Note $E_{C(INI)}$ is the energy required to initially charge $C_{VAR}$.

The challenge with keeping $q_C$ constant is $v_C$ can reach 100–300 V [32]–[33], well above the breakdown voltages of high-volume, low-cost semiconductor technologies. More expensive and specialized technologies, such as silicon-on-insulator (SOI) processes [33], can sustain these extreme voltages, but their increased costs may limit the extent to which the market will adopt them. Alternatively, keeping $V_{C(INI)}$ low enough to ensure $V_{C(FIN)}$ stays below, for example, in-package battery $V_{BAT}$ (e.g., 2.7–4.2 V) reduces $E_{GAIN}$ to

$$E_{GAIN} = \frac{0.5 C_{MIN}}{C_{MAX}} \Delta C_{VAR} V_{BAT}^2.$$  \hspace{1cm} (9)

Constraining $q_C$ this way is compatible with low-cost, high-volume processes, but at the expense of lower $E_{GAIN}$ (because $C_{MIN}/C_{MAX}$ is small). What is more, transferring energy to and from $C_{VAR}$ requires circuitry that dissipates additional power.

By holding $v_C$ rather than $q_C$, the mechanical energy that separates $C_{VAR}$’s plates decreases capacitance, which drives charge $q_C$ out of $C_{VAR}$ in the form of harvesting current $i_{HARV}$:

$$i_{HARV} = \frac{\partial q_C}{\partial t} = C_{VAR} \frac{\partial v_C}{\partial t} + V_{CVAR} \frac{\partial C_{VAR}}{\partial t} + V_C \frac{\partial C_{VAR}}{\partial t}.$$  \hspace{1cm} (10)

Although constraining $v_C$ this way is compatible with standard processes, typical implementations use an additional device (e.g., voltage source, capacitor, electret, etc.) to fix $v_C$ [34]–[37], which contradicts the goals of integration, complicates the assembly process, and requires an energy-transferring circuit to charge the device with the harvested energy. Constraining $v_C$ with the already-existing system battery (that is to receive charge energy) enhances integration because it does not require an additional source [38]–[41].

Switched-Inductor Harvester: A viable energy-harvesting cycle for a battery-constrained electrostatic system, as Fig. 8 illustrates, starts by precharging $C_{VAR}$ to $V_{BAT}$ when $C_{VAR}$ peaks at $V_{MAX}$. Afterwards, connecting $C_{VAR}$ to $V_{BAT}$ as $C_{VAR}$ falls to $V_{MIN}$ steers charge (harvests) from $C_{VAR}$ to $V_{BAT}$. The system then disconnects $C_{VAR}$ and allows its voltage to drop (reset) as $C_{VAR}$ rises again to $V_{MAX}$, at which point the cycle repeats. Notice that keeping $C_{VAR}$ attached to $V_{BAT}$ when $C_{VAR}$ rises would draw reverse (discharge) current from $V_{BAT}$.

During precharge, energizing $C_{VAR}$ at $V_{MAX}$ to $V_{BAT}$ represents an energy investment $E_{INV}$ from $V_{BAT}$ equivalent to

$$E_{INV} = 0.5 C_{MAX} V_{BAT}^2.$$  \hspace{1cm} (11)

Therefore, to produce a net energy gain $E_{NET}$, the energy harvested $E_{HARV}$ when subsequently connecting $C_{VAR}$ to $V_{BAT}$ (from $i_{HARV}$ or $V_{BAT} \Delta C_{VAR}/\Delta t$, where $\Delta C_{VAR}$ is $C_{MAX} - C_{MIN}$) must exceed $E_{INV}$ and all other losses in the system:

$$E_{HARV} = V_{BAT} i_{HARV}(t) dt = \Delta C_{VAR} V_{BAT}^2$$  \hspace{1cm} (12)

and

$$E_{NET} = E_{HARV} - E_{INV} - E_{LOSSES} = (0.5 C_{MAX} - C_{MIN}) V_{BAT}^2 - E_{LOSSES}.$$  \hspace{1cm} (13)

Fig. 8. Battery-constrained, switched-inductor electrostatic harvesting cycle.

When considering energy-transfer strategies to precharge $C_{VAR}$ to $V_{BAT}$, the most straightforward embodiment is to connect them with a switch. With this approach, however, the switch dissipates considerable power because it conducts current while dropping a considerable voltage: the initial difference between $V_{BAT}$ and $V_C$. More specifically, charging $C_{VAR}$ at $V_{MAX}$ from $V_{C(INI)}$ to $V_{BAT}$ by closing a switch demands $i_{BAT}$ from $V_{BAT}$, where

$$i_{BAT} = \left(\frac{V_{BAT} - V_{C(INI)}}{R_{SW}}\right) \exp\left(-\frac{t}{R_{SW} C_{MAX}}\right)$$  \hspace{1cm} (14)
and \[ V_c = V_{BAT} - i_{SW}R_{SW}, \] (15) which means the switch, irrespective of \( R_{SW} \), dissipates as much energy (\( E_{SW} \)) as \( C_{VAR} \) requires during precharge:

\[ E_{SW} = \int_0^{i_{BAT}} (V_{BAT} - V_c) \, dt = 0.5C_{MAX}(V_{BAT} - V_{C(INI)})^2 = E_{INV}. \] (16)

In other words, \( V_{BAT} \) must invest twice \( E_{INV} \) to charge \( C_{VAR} \) from \( V_{C(INI)} \) to \( V_{BAT} \), the transfer efficiency of which is 50%.

Transferring energy with an inductor (\( L_X \)), on the other hand, is virtually lossless because only a diminutive fraction of the voltage dropped appears across the interconnecting switches. In this way, the circuit in Fig. 9a transfers energy into \( L_X \) from the battery (\( V_{BAT} \)) to charge \( C_{VAR} \) to \( V_{BAT} \). Here, \( V_{BAT} \) first energizes both \( L_X \) and \( C_{VAR} \) by closing \( S_E \) through energizing time \( \tau_E \). Then, \( S_E \) opens and \( S_D \) closes to de-energize \( L_X \) fully into \( C_{VAR} \). \( S_D \) disengages when \( V_C \) reaches \( V_{BAT} \), marking the end of the precharge phase. The system includes a dead time between \( S_E \) opening and \( S_D \) closing to prevent shoot-through current from discharging \( V_{BAT} \).

![Fig. 9. (a) Switched-inductor power stage and (b) simulated waveforms.](image)

Because this energy-transfer process only lasts a small fraction of the vibration cycle, \( C_{VAR} \) remains virtually constant (at around \( C_{MAX} \)) through the precharge phase. \( S_E \) and \( S_D \) remain off during subsequent phases of the harvesting cycle to avoid discharging \( C_{VAR} \). The system then connects \( C_{VAR} \) to \( V_{BAT} \) with \( S_H \) to start the harvesting phase. As a result, as vibrations reduce \( C_{VAR} \) from 391 to 100 pF, for instance, as Fig. 9b demonstrates between 23.7 and 24.05 ms, \( i_{HARV} \) flows into the battery. The energy the battery accumulates in one cycle is sufficiently high to overcome its initial investment \( E_{INV} \) (at 2.75 nJ in Fig. 9b) plus the parasitic losses in the system (\( E_{LOSS} \)) to net a gain, in this case, of 1 nJ per cycle.

**C. Synchronization and Control**

Fig. 10 demonstrates a circuit embodiment of the electrostatic harvester. As in the piezoelectric case, all circuit blocks, except for \( L_X \), \( V_{BAT} \), and transducer \( C_{VAR} \), are on chip (where \( C_{VAR} \) can viably be on chip when integrated with a MEMS process). Similar to a buck converter, \( M_{PE} \) and \( M_{ND} \) implement \( L_X \)'s energizing and de-energizing switches \( S_E \) and \( S_D \), respectively. Harvesting switch \( S_H \) consists of back-to-back PMOS transistors \( M_{PHA} \) and \( M_{PHB} \) to ensure their parasitic body diodes do not engage during the precharge phase. All comparators only operate during their respective phases to conserve energy. Also, because vibrations are typically slow, sensing comparators \( CP_{P-STR} \) and \( CP_{H-END} \), which prompt the system to start and end the precharge and harvesting phases, respectively, need not respond quickly, which means they can operate in sub-threshold (with nA’s).

**Reset:** To precharge \( C_{VAR} \) to \( V_{BAT} \) at \( C_{MAX} \) and subsequently connect \( C_{VAR} \) to \( V_{BAT} \), the harvester must first monitor \( C_{VAR} \)'s capacitance. Fortunately, floating \( C_{VAR} \) in the reset phase, when capacitance rises to \( C_{MAX} \), induces \( V_C \) to fall from \( V_{BAT} \) to \( V_{BAT}C_{MIN}/C_{MAX} \) (in charge-constrained fashion). Sensing, as a result, when \( V_C \) reaches its minimum voltage indicates when \( C_{VAR} \) peaks. To this end, comparator \( CP_{P-STR} \) in Fig. 10 senses when \( V_C \) begins to rise above its delayed counterpart \( V_D \) to prompt the logic to start the precharge process.

![Fig. 10. Switched-inductor, voltage-constrained electrostatic harvester circuit.](image)

**Precharge:** The system must determine how long (\( \tau_E \)) \( L_X \) should energize to precharge \( C_{VAR} \) to \( V_{BAT} \). Charging \( C_{VAR} \) to \( V_{BAT} \) as close as possible is critical because, with an undercharged \( C_{VAR} \), \( S_H \) first discharges \( V_{BAT} \) to \( C_{VAR} \) before \( C_{VAR} \) can generate \( i_{HARV} \), which represents an additional loss in the system. Overcharging \( C_{VAR} \) also incurs a loss because \( S_H \) would first discharge \( C_{VAR} \) to \( V_{BAT} \) before \( C_{VAR} \) could generate \( i_{HARV} \). The harvester in Fig. 10 therefore uses a tuned \( \tau_E \) to precharge \( V_C \) to \( V_{BAT} \) precisely. A dead time after \( \tau_E \), \( CP_{P-END} \) detects when \( L_X \) depletes (i.e., \( i_L \) drops to zero) by sensing when switching node \( V_{SW} \) decreases to 0 V. \( CP_{P-END} \) then triggers \( S_H \)'s S-R latch to start the harvesting phase.

Instead of tuning \( \tau_E \) externally, the circuit in Fig. 11 implements a feedback loop that automatically adjusts \( \tau_E \) to the optimal length [41]. The loop essentially samples and compares \( C_{VAR} \)'s voltage at the end of the precharge phase (\( V_{CEND} \)) to \( V_{BAT} \) and incrementally adjusts \( \tau_E \) on a cycle-by-cycle basis until \( V_{CEND} \) finally reaches \( V_{BAT} \) accurately. More specifically, \( CP_{VC} \) terminates \( L_X \)'s energizing time \( \tau_E \) when \( C_{VAR} \) charges to intermediate voltage \( V_{REF} \). Then, \( CP_{REF} \) compares \( V_{CEND} \) and \( V_{BAT} \) to determine whether to increment \( V_{REF} \) (and therefore \( \tau_E \)) up or down until \( V_{CEND} \) nears \( V_{BAT} \).

![Fig. 11. Feedback loop for automatically tuning inductor energizing time \( \tau_E \).](image)

Neglecting parasitic losses in the system, \( V_C \) should charge to \( 0.5V_{BAT} \) (i.e., \( V_{REF} \) is 0.5\( V_{BAT} \)) during \( \tau_E \) to end at \( V_{BAT} \) after the system exhausts \( L_X \) into \( C_{VAR} \) [40]. In practice, however, the circuit dissipates power so the system must draw more energy (than theoretical \( E_{INV} \)) from \( V_{BAT} \) to compensate for this loss, which means \( V_{REF} \) actually exceeds 0.5\( V_{BAT} \). Also, since precharge lasts for a diminutive fraction of the entire
cycle (e.g., about 200 ns of 33 ms), circuits in the precharge block must react quickly, which means they should operate in strong inversion. Although strong-inversion currents (i.e., power) are not small, their corresponding energy losses are because current only flows for a short while.

**Harvest:** In the harvesting phase, \( i_{\text{HARV}} \) flows through \( S_H \)'s turn-on resistance to raise \( V_C \) slightly above \( V_{\text{BAT}} \), keeping \( C_{\text{PH-END}} \)'s output from resetting \( S_H \)'s S-R latch. Once \( C_{\text{VAR}} \) reaches \( C_{\text{MIN}} \) and \( i_{\text{HARV}} \) consequently falls to zero, \( V_C \) drops to \( V_{\text{BAT}} \) and \( C_{\text{PH-END}} \) trips to end the harvesting phase, resetting the latch and disengaging \( S_H \). If the system undercharges \( C_{\text{VAR}} \) in the precharge phase, \( C_{\text{PH-END}} \) can trip ahead of time, possibly triggering an additional precharge event. For this reason, the harvesting phase must end when \( V_C \) falls (not rises) from above \( V_{\text{BAT}} \) to \( V_{\text{BAT}} \), which is why the decision should rest on \( C_{\text{PH-END}} \)'s rising or falling edge.

IV. **ENERGY LOSSES IN SWITCHED-INDUCTOR HARVESTERS**

The fundamental advantage of switched-inductor circuits in piezoelectric and electrostatic harvesters is efficiency, because switches conduct current when their terminal voltages are close to 0 V. While efficiencies are considerably high, conduction losses across equivalent series resistances (ESR) (in switches, inductor, and capacitors) and body diodes, switching losses from charging and discharging parasitic gate capacitances (in power switches and drivers), and bias/quietest power (in the controller) keep efficiencies from nearing 100%. Inductor current \( i_L \), for example, dissipates conduction power \( P_{C(R)} \) across relevant ESRs. Because \( i_L \) rises and falls linearly back to zero in every cycle, these converters operate in DCM, which means losses per cycle \( E_{C(R)} \) are

\[
E_{C(R)} = N \left( P_{C(R)} \tau_C \right) = \left( \frac{I_{\text{L}(PK)}}{\sqrt{3}} \right)^2 R_{\text{QD}} \tau_C, \tag{17}
\]

where \( N \) is the number of times an inductor transfers energy within one vibration cycle (e.g., two and one for the piezoelectric and electrostatic cases presented), \( I_{\text{L}(PK)} \) is \( i_L \)'s peak, \( R_{\text{EQ}} \) is the combined ESR of the conduction path, and \( \tau_C \) is the inductor's conduction time. During the dead time \( \tau_{\text{DT}} \) of interconnecting switches, \( i_L \) also dissipates energy \( E_{C(D)} \) across conducting body diodes:

\[
E_{C(D)} = N \left( I_{\text{L}(PK)} \right)^2 R_{\text{EQBDD}} \tau_{\text{DT}}, \tag{18}
\]

where \( R_{\text{EQBDD}} \) is the combined ESR of the conduction path and \( V_{\text{BD}} \) the diode's voltage drop. (Note the self-synchronized MOSFET in Fig. 7 minimizes \( \tau_{\text{DT}} \) and the loss it causes.)

During certain switching events, some transistors also dissipate I-V overlap power because they conduct current while their drain-source voltages \( V_{\text{DS}} \) concurrently transition several Volts. The electrostatic power-stage in Fig. 10, for example, starts the precharge phase by engaging \( M_{\text{HE}} \), except \( i_L \) is zero when that happens so \( M_{\text{HE}} \) dissipates little power. When \( M_{\text{HE}} \) shuts off, however, \( i_L \) peaks and \( M_{\text{HE}} \) carries \( i_L \) while \( V_{\text{SW}} \) (and \( V_{\text{SD}} \)) transitions between \( V_{\text{BAT}} \) and a diode voltage \( (V_D) \) below ground, until \( M_{\text{ND}} \)'s body diode conducts \( i_L \). When \( M_{\text{HE}} \)'s drain-gate capacitance sets overlap time \( \tau_{\text{OV}} \), \( M_{\text{HE}} \)'s average \( V_{\text{SD}} \) is 0.5\((V_{\text{BAT}} + V_D)\) so \( E_{\text{SW(IV)}} \) is roughly

\[
E_{\text{SW(IV)}} = 0.5(V_{\text{BAT}} + V_D) \tau_{\text{OV}} \tag{19}
\]

\( M_{\text{ND}} \), on the other hand, only transitions one diode voltage when engaging and \( i_L \) nears zero when disengaging, which means \( M_{\text{ND}} \) dissipates negligible overlap losses. Drivers also dissipate power to charge and discharge gate capacitances:

\[
E_{\text{SW(GD)}} = \sum \left( C_{\text{GIR}} V_{\text{DRV}} \right)^2, \tag{20}
\]

where \( E_{\text{SW(GD)}} \) is gate-drive energy per cycle, \( C_{\text{GIR}} \) is equivalent gate capacitance, \( V_{\text{DRV}} \) is gate-drive voltage, and all switches (in Figs. 6 and 10) engage and disengage only once per cycle.

The controller, which generates the switching signals necessary to condition and synchronize the system to vibrations, also dissipates quiescent (bias) power. The circuits comprise it (e.g., in Figs. 6, 10, and 11) generally fall into four categories: fast comparators, slow comparators, logic, and bias. Comparators that monitor \( i_L \) (e.g., \( CP_{\text{HE}}, CP_{\text{DN}}, CP_{\text{P-END}}, \) and \( CP_{\text{VC}} \) from Figs. 6 and 10), for example, must be quick because they transfer energy in sub-\( \mu \)s of the ms period; monitoring when \( C_{\text{VAR}} \)'s voltage peaks (e.g., \( CP_{\text{REF}} \) in Fig. 11) also occurs in sub-\( \mu \)s. Although these comparators require considerable strong-inversion power, they only operate for a diminutive fraction of the period, so they dissipate little energy. Comparators that monitor the transducer (e.g., \( CP_{\text{PK}}, CP_{\text{P-START}}, \) and \( CP_{\text{H-END}} \)) in Figs. 6 and 10) need not respond in \( \mu \)s because they monitor ms events, so they can be slow. Unfortunately, they, like bias circuits, operate through the vibration cycle, which means they can dissipate considerable energy, if not biased in sub-threshold (with nA's). Logic consumes relatively little energy because the frequency of vibrations is low (e.g., 50 – 300 Hz) and transistors are small.

Since conduction losses increase with channel resistance and switching losses with gate capacitance, smaller transistors dissipate more conduction power and less switching losses than larger ones, and vice versa. Accordingly, using minimum channel lengths and balancing transistor widths in the power stage (of Figs. 5a and 9a) are important to minimize overall losses in the system. Another design tradeoff is the number of times the inductor energizes and de-energizes in one cycle. Transferring energy multiple times, for example, increases the number of switching events (and associated switching losses) and decreases \( i_L \)'s peak (which reduces conduction power). Also, since piezoelectric power is proportional to the square of the peak voltage, harnessing energy at the positive and negative peaks (i.e., twice per cycle) is optimal.

V. **EXPERIMENTAL VALIDATION**

The shaker in Fig. 12 generated the periodic vibrations that validated the functionality of the circuits prototyped. The piezoelectric harvester tested used a 4.4 × 4.4 × 1.4-mm\(^3\), 160- \( \mu \)H inductor with ESR of 3.4 \( \Omega \) from Coilcraft and a 44 × 13 × 0.4-mm\(^3\) piezoelectric cantilever from PiezoSystems, which with 275 nF and 10 M\( \Omega \) of parasitic capacitance and resistance, resonated at about 100 Hz. The electrostatic case employed a 2 × 2 × 1-mm\(^3\), 10- \( \mu \)H inductor with ESR of 1 \( \Omega \) from Coilcraft and a 10.16 × 20.32 × 0.172-cm\(^3\) variable capacitor prototype built in the laboratory (to emulate the
operation of a MEMS counterpart). At 30 Hz, the capacitor oscillated between 157 and 991 pF. A low-leakage, unity-gain op amp then monitored the voltage of the battery or capacitor that these harvesters ultimately charged.

With this setup, the prototyped piezoelectric harvester charged 23 µF from 2.6 to 4 V in 28.5 sec. from vibrations at 100 Hz, as Fig. 13a illustrates, which represents 3.73 µW of average power, 37.3 nJ/cycle, and 106.3 µJ of total energy. In harvesting, the system increased peak piezoelectric voltages \(v_{\text{PZT}}\) from 0.46 (when disconnected –not shown–) to 0.65 V (when connected, as shown), which indicates the harvester conditioned the transducer to induce it to generate more power. Similarly, Fig. 13b shows the operation of the prototyped electrostatic harvester charging 1 µF from 2.7 to 4.2 V in 68.8 sec. from vibrations at 30 Hz, which equates to 75.2 nW of average power, 1.930 and 3.885 nJ/cycle at 2.7 and 4.2 V, and 5.12 µJ of total energy.

When charging a 3.5-V battery, the prototyped electrostatic harvester drew (on average) up to 505 nA from \(C_{\text{VAR}}\) [40]–[41]. The 10.1 nJ/cycle gained was sufficient to overcome \(V_{\text{BAT}}\)'s initial 6.72 nJ/cycle precharge investment in \(C_{\text{VAR}}\). Since 991 pF requires 6.07 nJ to precharge to 3.5 V, the switched-inductor precharger’s efficiency was 90.3%. The prototype kept lost quiescent energy \(E_0\) in the controller low at 0.91 nJ/cycle by operating fast, power-hungry blocks only on-demand and their slower counterparts in sub-threshold with nA’s. After subtracting the initial investment and all other losses in the system (from Table I), the net energy gain per cycle was 2.47 nJ, which is equivalent to 74.1 nW at 30 Hz.

### Table I. Measured Energy Per Cycle in the Electrostatic Harvester Prototype.

<table>
<thead>
<tr>
<th>Phase</th>
<th>Component</th>
<th>(i_0)</th>
<th>(\tau_0)</th>
<th>(E_o)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Reset</td>
<td>(C_{\text{PSTART}})</td>
<td>2.0 nA</td>
<td>13.3 ms</td>
<td>0.095 nJ</td>
</tr>
<tr>
<td></td>
<td>(C_{\text{PVC}})</td>
<td>28.0 µA</td>
<td>230 ns</td>
<td>0.023 nJ</td>
</tr>
<tr>
<td>Precharge</td>
<td>(C_{\text{PEND}})</td>
<td>22.4 µA</td>
<td>253 ns</td>
<td>0.020 nJ</td>
</tr>
<tr>
<td>Local µA-Bias</td>
<td>12.4 µA</td>
<td>500 ns</td>
<td>0.022 nJ</td>
<td></td>
</tr>
<tr>
<td>Harvest</td>
<td>(C_{\text{PREF}})</td>
<td>3.0 nA</td>
<td>20 ms</td>
<td>0.21 nJ</td>
</tr>
<tr>
<td>Optimal (\tau_c)</td>
<td>(C_{\text{REF}})</td>
<td>42.3 µA</td>
<td>489 ns</td>
<td>0.072 nJ</td>
</tr>
<tr>
<td>Tuning</td>
<td>Variable</td>
<td>34.0 µA</td>
<td>489 ns</td>
<td>0.058 nJ</td>
</tr>
<tr>
<td>Reference</td>
<td>nA-Bias</td>
<td>100 nA</td>
<td>28.0 µA</td>
<td>0.036 nJ</td>
</tr>
</tbody>
</table>

The prototyped piezoelectric harvester drew 1.6 to 29.6 µW [42]–[43] from periodic vibrations with accelerations at the base of the cantilever that ranged from 0.05 to 0.16 m/s² – note typical HVAC vents in the office exhibit accelerations that exceed 0.2 m/s² [8]. Efficiency remained below 50%, as Fig. 14a shows, because conduction losses \(P_c\) dominate over their switching and quiescent counterparts \(P_{SW}\) and \(P_Q\). Improving efficiency here is possible by decreasing resistances (i.e., power losses) in the conduction path, that is, by balancing losses. In other words, enlarging the switches and using a physically larger inductor, which has lower equivalent series resistance (ESR), increases efficiency, as the simulated results in Fig. 14b demonstrate (with switches that are \(10\times\) larger and a 0.84-Ω 160-µH inductor that is \(5 \times 5 \times 3\) mm³). The tradeoff, of course, is space (i.e., cost) because both die size and board space increase. Nevertheless, with these values, \(P_c\) decreases more than gate-drive losses \((P_{SW})\) increase, and efficiency, as a result, reaches 75%. The design, however, is now optimum for a higher power range, which means efficiency drops below 50% at lower power levels, when \(P_{SW}\) and \(P_Q\) dominate. Plus, increasing the power range beyond the balanced region again induces \(P_c\) to dominate and efficiency to suffer.

![Fig. 12. Setup used to evaluate kinetic piezoelectric/electrostatic harvesters.](Image)

![Fig. 13. Experimental (a) piezoelectric and (b) electrostatic charging profiles.](Image)

![Fig. 14. Efficiency and losses in the piezoelectric harvester with physically (a) smaller (experimental) and (b) larger (simulated) switches and inductor.](Image)
mechanical domain) and (ii) the energy-transfer process is shorter than the vibration period.

![Graph showing experimental (piezo-) charging profile from non-periodic vibrations.](image)

**Fig. 15.** Experimental (piezo-) charging profile from non-periodic vibrations.

**VI. CONCLUSION**

Since energy per cycle is low in miniaturized systems and synchronizing a circuit to vibrations and transferring energy dissipate power, the fundamental challenge in harvesting kinetic energy is producing a net energy gain, in other words, reducing losses. This paper presents a tutorial on how switched-inductor converters can not only diminish losses in energy-harvesting systems but also condition the transducer's voltage to expand the harvestable vibration range and therefore induce the transducer to draw more energy from vibrations. Inductors, however, are bulky and difficult to integrate, which is why using only one is so important. Still, tiny transducers generate little power and may therefore lose a considerable portion to conduction and switching losses, reducing efficiencies to 40%. When optimally designing the switches, inductor, and number of switching events for the power level targeted, efficiency performance can potentially surpass 90%, even when transferring only 6.7 nJ per cycle. To reduce quiescent losses to less than 1 nJ per cycle, the slow portions of the controller should operate in sub-threshold with ultra-low currents and fast blocks, which require more current, operate for a diminishing fraction of the vibration period. After all, continuously generating output power, even if only a few nW's or μW's, can charge a battery so, when a sensor needs energy, which typically does not happen often, the battery can readily supply it. The idea is to supplement the system with enough energy over time to extend its operational life and avoid having to replace an exhaustable battery.

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