AN ELECTROSTATIC CMOS/BICMOS LI ION VIBRATION-BASED HARVESTER-CHARGER IC

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AN ELECTROSTATIC CMOS/BICMOS LITHIUM VIBRATION-BASED HARVESTER-CHARGER IC

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To my wife Johanne and my son Yonel Omar
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SUMMARY

Self-powered microsystems, such as wireless transceiver microsensors, appeal to an expanding application space in monitoring, control, and diagnosis for commercial, industrial, military, space, and biomedical products. As these devices continue to shrink, their microscale dimensions allow them to be unobtrusive and economical, with the potential to operate from typically unreachable environments and, in wireless network applications, deploy numerous distributed sensing nodes simultaneously. Extended operational life, however, is difficult to achieve since their limited volume space constrains the stored energy available, even with state-of-the-art technologies, such as thin-film lithium-ion batteries (Li Ion) and micro-fuel cells. Harvesting ambient energy overcomes this deficit by continually replenishing the energy reservoir and, as a result, indefinitely extending system lifetime.

In this work, an electrostatic harvester that harnesses ambient kinetic energy from vibrations to charge an energy-storage device (e.g., a battery) is investigated, developed, and evaluated. The proposed harvester charges and holds the voltage across a vibration-sensitive variable capacitor so that vibrations can induce it to generate current into the battery when capacitance decreases (as its plates separate). The challenge is that energy is harnessed at relatively slow rates, producing low output power, and the electronics required to transfer it to charge a battery can easily demand more than the power produced. To this end, the system reduces losses by time-managing and biasing its circuits to operate only when needed and with just enough energy while charging the capacitor through an efficient quasi-lossless inductor-based precharger. As result, the
proposed energy harvester stores a net energy gain in the battery during every vibration cycle.

Two energy-harvesting integrated circuits (IC) were analyzed, designed, developed, and validated using a 0.7-µm BiCMOS process and a 30-Hz mechanical variable capacitor. The precharger, harvester, monitoring, and control microelectronics of the first prototype draw sufficient power to operate and at the same time produce experimentally 1.27, 2.14, and 2.87 nJ per vibration cycle for battery voltages at 2.7, 3.5, and 4.2 V, which with 30-Hz vibrations produce 38.1, 64.2, and 86.1 nW. By incorporating into the system a self-tuning loop that adapts optimally the inductor-based precharger to varying battery voltages, the second prototype harnessed and gained 1.93, 2.43, and 3.89 nJ per vibration cycle at battery voltages 2.7, 3.5, and 4.2 V, generating 57.89, 73.02, and 116.55 nW at 30 Hz. The harvester ultimately charges from 2.7 to 4.2 V a 1-µF capacitor (which emulates a small thin-film Li Ion) in approximately 69 s, harnessing in the same length of time 47.9 % more energy than with a non-adapting harvester.
CHAPTER 1

SELF-POWERED MICROSYSTEMS

Modern state-of-the-art microelectronic technologies enable ultra-compact microscale devices that provide monitoring and sensing functions to a wide variety of applications where their minuscule dimensions allow them to be unobtrusive and economical, with the potential to operate from difficult-to-reach environments and, in wireless network applications, deploy numerous distributed sensing nodes simultaneously [1]-[6]. Some examples, to name a few, include microsensors that monitor environmental conditions like temperature, humidity, and air pressure from remote or inaccessible areas [1], [7]-[10], and diagnose, from within, the health of industrial machinery (e.g., motors, pumps, and turbines) [11]-[12], aircraft [13]-[14], space satellites [15], electromagnetic interference (EMI) shielding [16], and large-scale structures like concrete bridges [17]. In the biomedical field [18]-[20], low-profile implantable or wearable microsystems supervise vital signs of patients (such as blood pressure, heartbeats, and exertion levels) [12], [21]-[25] and detect DNA sequences, viruses, or other disease markers [26]. They can also deliver drugs (e.g., insulin for diabetic patients) internally [27], supervise local on-site radiation levels on cancer patients [28], and stimulate specific muscle and neural tissues [29]-[30]. For each of these examples, however, maintenance and upkeep of each microsystem is impractical due to their diminutive size, large distribution in a network, and inaccessibility. For this reason, each microsensor node must be autonomous and self-powered from an in-package energy supply. As a result, the initial amount of energy
stored, which is limited by the lack of available space, dictates the operational life of each device.

To prolong their life, self-powered microsystems must perform at a high efficiency to make the most of the restricted available energy. This means that each component of the system, which, as shown in Figure 1.1, typically includes sensor and interface circuits, digital data conversion and storage, transceiver, and power management, must dissipate the least amount of energy possible [31]-[33], and, in some instances, even sacrificing performance (e.g., speed and accuracy) for the sake of efficiency [31]. To this end, data transmission, which is the predominant energy consumer of the system, occurs at low rates (less than 10 kb/s) and broadcasts to short distances (less than 10 m) [16]. Data is modulated using simple schemes like frequency shift keying (FSK) that simply switches between discrete carrier frequencies and, in low-interference environments, on-off keying (OOK) that turns on or off the carrier signal [16]. To save energy, the system does not condition the sensed signal, but converts it to a digital word through an analog-to-digital converter (ADC) and stores it in memory until it is transmitted to a base station. What is more, if in a large wireless network, each sensor node transmits data using energy-aware protocols and algorithms to reach find the most

![Figure 1.1 Typical wireless microsensor architecture.](image-url)
efficient transmission path to the base station [1], [32]-[35].

Fortunately, since typical applications respond only to sporadic events or slow-changing variables like temperature and air pressure, the main energy-saving strategy is to duty-cycle power-intensive tasks such as wireless transmission and sensing [16], [36]-[37]. These functions do not engage continuously or simultaneously, but periodically or on-demand, allowing the system to idle often by entering deep-sleep modes where most of the system shuts off and drain energy at a slower rate (or in other words, operate at a lower average power). By offsetting each task to turn on for short intervals (0.001-5 %) of the entire period, the system’s average power consumption drops significantly. As an example, wireless transmission and reception can each consume 5 mW and 3.75 mW, respectively, for 5 ms, while a typical sensing function dissipates 10 µW for 1 ms [36]. Enabling each task once every 500 ms period (2.2% duty cycle) lowers the average power consumption of the system to 87.52 µW, as Figure 1.2 illustrates. For even slower sensing applications, waiting instead 500 s (0.002% duty cycle) between load events further reduces the average power consumption to 87.52 nW. In this way, by sporadically

![Figure 1.2 Load profile of a duty-cycled microsystem.](image-url)
waking up only when necessary and slowing down its energy use, the system extends its life.

Because of severe space and volume constraints, incorporating high-energy density supplies, such as nuclear batteries and micro-fuel cells, are critical to enable microsystems to lengthen their functional survival. Nuclear microbatteries, for instance, draw energy from beta particles (high-energy electrons) emitted from decaying radioactive material, such as nickel-63 or tritium, to generate substantial energy densities around 850 kWhr/kg [37]-[40]. However, the use of radioactive materials raises safety and cost concerns. On the other hand, direct-methanol (DM) proton-exchange membrane (PEM) fuel cells still offer relatively high energy densities in the 1 kWhr/kg range, which, although not as great as nuclear batteries, produce safer byproducts (carbon dioxide and water). What is more, microelectromechanical systems (MEMS) technologies permit microscale integration [41]-[46], even alongside complimentary metal-oxide semiconductor (CMOS) electronics [47]. However, slow reaction times (i.e., low power levels near 10 W/kg), dreadful output voltages (0.4-0.7 V), and fuel leakages limit their use. Ultimately, either nuclear batteries or fuel cells store a finite amount of energy that is bound to deplete eventually, regardless of how efficient the system is or how low its average consumption power is. By extracting energy from its surroundings, however, energy harvesters can replenish the system’s consumption and extend its life indefinitely, independent of the initial amount of energy stored on-board [48].

Even though duty-cycling high-power tasks helps lower the average power consumption of the system, it does little to attenuate its high peak power requirements. This presents a challenge because even though harvesters extract energy from a relatively
endless source in the environment, they do so at slow rates (and therefore low power levels) and cannot provide the required high peak power bursts. What is more, ambient energy can be variable and unpredictable and might not be available at all times and therefore not supplied continuously. For these reason, a more practical approach is to store, when available and possible, converted energy in devices capable of supplying power on-demand, when needed [48]. This approach complements the quasi-infinite energy sourcing capabilities of harvesters with a high-power source that simultaneously supplies power-demanding functions and stores the harnessed ambient energy, as the Ragone plot in Figure 1.3 shows.

Passive energy-storage elements, such as capacitors and inductors, store energy in either electric or magnetic fields, respectively, which can be released quickly leading to substantial power densities. Capacitors, for example, can supply instantaneous current, but can only sustain it for short bursts because of their low energy storage capabilities. Ultracapacitors, also known as electrochemical double-layer capacitors, store significantly more energy (2-10 Whr/kg) due to considerably greater capacitances (~1
mF/mm$^3$) created by the microscopic separation of ions that spread across porous carbon-based electrodes [37], [49]-[51]. However, they suffer from high costs, considerable current leakage (in the range of 1-20 µA), which contradicts the efficiency requirements of self-powered devices, and microscale integration difficulties [37], [52]. Inductors also exhibit low energy densities when integrated in a microsystem because of small inductance values and weak magnetic fields due to few coil turns and lack of strong magnetic core (20-40 nH/mm$^2$) [53]. Additionally, inductors cannot store energy for long periods and must either receive, release, or lose the energy. In conclusion, capacitors and inductors are not capable of sustaining high-power tasks for long, but are useful and suitable as intermediate short-term energy transfer devices.

Electrochemical batteries store electrical energy chemically, which is released through a reaction inside the battery cell that transfers electrons from its anode to its cathode across an electrolyte material [54]. Recharging a secondary (i.e., a rechargeable) battery reverses this reaction and stores electrical energy back in the form of chemical bonds. The chemical reaction releases energy at a slower rate than passives can, and, therefore, batteries exhibit lower power densities, but at significantly higher energy levels capable of sustaining wireless transmission loads. Conventional chemistries, such as nickel-metal hydride (NiMH) and nickel cadmium (NiCd), offer good power densities and discharge rates, but also feature short cycle life and adverse memory effects [54]. Lithium-ion batteries (Li Ion), which are based on the exchange of lithium ions (Li$^+$) between electrodes, overcome these drawbacks with superior energy and power density, discharge rate, cell voltage, longer cycle life, and nonexistent memory effects. More importantly, thin-film technology (less than 15 µm thickness) permits the integration of
lithium-ion batteries into a microsystem while capable of delivering relatively high power levels (~200 W/kg) with modest energy levels (200-400 Whr/kg) that can sustain power-intensive tasks, such as wireless telemetry [55]-[57]. Thin-film lithium-ion batteries feature capacities in the range of 100-200 μAhr/cm\(^2\) while delivering up to 3 mA/cm\(^2\) by depositing cathode, anode, and electrolyte materials onto silicon-based integrated circuits (IC) using low-temperature physical vapor deposition techniques, such as sputtering and thermal evaporation [57]-[59]. A disadvantage is their sensitivity to overcharge and overdischarge. If the cell voltage increases above 4.2 V, or decreases below 2.7 V, the battery will significantly degrade, and in some cases, even vent and explode, but can be avoided with careful management [54].

A 2.5-mg thin-film Li Ion could viably hold up to 1 mWhr (3.6 J) inside a 1-mm\(^3\) (1-μL) volume, covering approximately 8.2 x 8.2 mm\(^2\) with 15-μm thin films [57]. If the system loads, as shown in Figure 1.2, are duty-cycled to an average power consumption of 87.52 μW (500-ms duty cycle period), the thin-film Li Ion can only sustain operations for 11.4 hours (without considering other losses such as self-discharge). Further increasing the duty cycle period by several orders of magnitude to 500 s can reduce average power consumption to 87.52 nW, sufficient to extend the life of the self-powered device to 476 days (more than a year). Nevertheless, its life is limited by the amount energy initially stored by the thin-film microbattery. The Li Ion’s high power density and integration potential, however, warrant its use to complement energy harvesters in a microsystem application, where the Li Ion represents an intermediary energy-storage device that accumulates scavenged energy while powering the system’s peak power loads.
Fundamentally, energy harvesters are extremely low power sources with a virtually boundless and inexhaustible energy supply (Figure 1.3) that, in the envisioned system, harness energy from the environment and channel it to charge a Li Ion, the high-power storage device. Harvesters alone may not be able to supply any load but they can refill an otherwise exhaustible reservoir of energy and extend system life. The harvester scavenges energy to replenish the system’s consumption continuously and restores it into the battery, which accumulates the energy during low-power sleep modes. Essentially, the battery behaves as a power cache, constantly receiving a charge trickle from the harvester and supplying seldom-occurring power-intensive functions on-demand, but only if sufficient energy is available. Therefore, by harnessing, converting, and transferring additional ambient energy to charge the Li Ion, self-powered and autonomous microdevices self-replenish and self-sustain to extend their lifetime almost indefinitely, independent of the initial energy stored on reservoir, barring the wear-and-tear effects of the components.
CHAPTER 2

AMBIENT ENERGY SOURCES

Present-day technologies can scavenge kinetic, thermal, light, and radio-frequency (RF) electromagnetic energy from the environment that, when compared to the energy stored in common storage elements like batteries and fuel cells, represents a relatively inexhaustible and infinite source of energy [60]-[64]. For this reason, energy harvesting methods must be characterized by their power density, rather than energy density, or in other words, by how fast energy can be converted. Table 2.1 compares the estimated power densities and challenges of various ambient energy sources based on previous work, but it is important to note that energy harvesting research is still in its infancy and lacks a standard for measurement of device performance. Consequently, published results feature different metrics that make a strict comparison between

<table>
<thead>
<tr>
<th>Energy Source</th>
<th>Challenge</th>
<th>Estimated Power (in 1 cm$^3$ or 1 cm$^2$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Light</td>
<td>Conform to small surface area</td>
<td>10 µW – 22.5 mW (Outdoors: 0.15 – 22.5 mW)</td>
</tr>
<tr>
<td></td>
<td></td>
<td>(Indoors: &lt;10 µW)</td>
</tr>
<tr>
<td>Vibrations</td>
<td>Variability of vibration</td>
<td>1 – 400 µW</td>
</tr>
<tr>
<td></td>
<td></td>
<td>(Piezoelectric: &lt; 400 µW)</td>
</tr>
<tr>
<td></td>
<td></td>
<td>(Electrostatic: &lt; 100 µW)</td>
</tr>
<tr>
<td></td>
<td></td>
<td>(Electromagnetic: &lt; 100 µW)</td>
</tr>
<tr>
<td>RF Energy</td>
<td>Power decreases drastically</td>
<td>0.01 - 0.1 µW (25-100 m from GSM station)</td>
</tr>
<tr>
<td></td>
<td>with distance</td>
<td></td>
</tr>
<tr>
<td>Thermal</td>
<td>Low thermal gradients</td>
<td>15 µW</td>
</tr>
<tr>
<td></td>
<td></td>
<td>(10 ºC gradient)</td>
</tr>
</tbody>
</table>
technologies difficult. More important, though, is that the power levels that can be achieved depend on the specific conditions surrounding the application and the compatibility of the available technologies. Light, for instance, can generate significant power levels if available, but it is highly dependent on how much exposure the device is subjected to (e.g., indoors versus outdoors), while thermal energy is limited to applications where significant temperature differences exist. Mobile sensors, on the other hand, may better reap the advantages of systematic vibrations and movements. As a result, it is unreasonable to compare each harvesting technology merely based on harvesting power density without considering the target application and the specific demands and considerations of the environment where it will deployed in.

2.1 Light Energy

Photovoltaic cells generate electrical energy from incident light [65]-[72]. Each cell consists of a pn+ junction, where light interfaces with the heavily doped and narrow n+-region, as shown in Figure 2.1(a). Photons with sufficient energy (that is, appropriate wavelength) create electron-hole pairs within the depletion region, where the junction built-in electric field immediately separates each so that electrons drift to the n+-side and holes to the p-side. Positive and negative charges accumulate at the p- and n+-side, respectively, developing an open-circuit voltage that is typically about 0.4-0.6 V. With a load connected across the cell, excess electrons travel through it from the n+-side to recombine with holes at the p-side, generating photocurrent $i_{PH}$ that is directly proportional to light intensity.
A current source models the generated photocurrent with a shunt parasitic diode representing the pn+ junction, as illustrated in Figure 2.1(b). The total current delivered to the load $i_{\text{Load}}$ is the difference between $i_{\text{PH}}$ and shunting diode current $i_D$:

$$i_{\text{LOAD}} = i_{\text{PH}} - i_D,$$

where the load voltage determines the forward bias diode current. As load voltage increases, more diode current shunts away photocurrent, and the total current flowing through the load decreases. The medium through which each generated carrier travels to reach the load introduces series resistance $R_S$, which is why, to minimize it, one of the electrodes completely covers the dark side, and numerous thin finger electrodes permeate the illuminated side without significantly impeding light exposure. Similarly, shunt resistance $R_P$ represents photogenerated carriers flowing through the surface and edges of the device rather than the load. Because of size constraints, a single photovoltaic cell cannot provide significant power and, consequently, several cells are connected in parallel arrays to increase current; conversely, cells are placed in series to augment their output voltage.
Photovoltaic cells are the most mature and commercially established energy-harvesting technology. Microscale systems typically employ thin-film photovoltaic cells, which convert available light energy at lower efficiencies (10-20 %) than their macroscale counterparts can [65]-[67]. The reasons efficiencies are lower are that up to 25 % of the available solar energy exists outside the spectrum range that silicon-based films can absorb, generated electron-hole pairs recombine near the crystal surface, and up to 20 % of the light reflects [65]. Even then, although power densities drop as dimensions shrink, crystalline and amorphous silicon-based thin-film cells harvest significantly greater power levels than other ambient sources. For instance, 0.35-μm CMOS photovoltaic diodes scavenge up to 22.5 mW/cm², with white light intensities similar to that of a sunny day outdoors, by using the p-type substrate and n-well diffusions as the anode and cathode of the diodes [73]. It is also possible to increase the effective internal surface area to improve efficiency, and therefore power density, by implementing a three-dimensional (3-D) diode structure geometry constructed on porous silicon [74]. Furthermore, maximum power point tracking (MPPT) strategies ensure that photovoltaic cells extract as much power as possible by dynamically adjusting the cell voltage [75]-[76]. Overall, energy from light potentially offers the highest output power levels, relative to other energy conversion mechanisms, in addition to being a mature IC-compatible technology. Nevertheless, its power output is heavily dependent on environmental conditions and changes drastically, by orders of magnitude, as light intensity varies.

The main limitation of harvesting energy from light is its availability. For instance, even if outdoors, energy varies with location, time, season, and atmospheric
conditions [76], not to mention that it is only available part of the day. Even in a sunny, cloudless day, light intensity changes as the day progresses because the angle of incidence varies. Indoors power levels drop drastically and photovoltaic cells generate 10 \( \mu \text{W/cm}^2 \) of usable electrical power under best-case conditions [60]. Consequently, the advantages that light as an energy source offers strongly depend on the conditions surrounding the microsystem, and even if available can fluctuate considerably.

2.2 Thermal Energy

Thermal gradients in the environment are directly converted to electrical energy through the Seebeck, or thermoelectric, effect [77]-[90]. Temperature differentials between opposite segments of a conducting material result in heat flow and, consequently, charge flow: hot carriers diffuse towards the cold end, while, conversely, cold carriers diffuse towards the hot end. As carriers diffuse, a concentration gradient produces an electric field across the material that opposes the net diffusion of carriers, eventually forcing equilibrium conditions. At equilibrium, the carriers traveling back to

![Figure 2.2](image)

Figure 2.2 (a) Thermoelectric energy converter based on two series thermocouples and (b) its circuit model.
the high temperature junction due to the electric field cancel the net number of hot
carriers diffusing towards the cold junction, and therefore, no voltage is established.

A thermocouple configuration in Figure 2.2(a) allows a more suitable approach
for power generation by electrically joining n- and p-type thermoelectric materials at the
hot end. Heat flow transports the dominant charge carriers of each material (electrons in
n-type and holes in p-type) to the low temperature junction, respectively ionizing each
base electrode with an opposite charge and therefore establishing a voltage differential.
As charge carriers depart the hot end, they leave behind ionized molecules that, instead of
attracting the opposite flow of charge from the material itself, attract carriers from the
opposite type material through a metallic, low impedance short. Electron carriers, for
instance, continually move from the p- to n-type material to de-ionize a molecule, at
which point they absorb thermal energy and continue to flow to the low temperature
junction. In the end, the generated voltage per couple, modeled by voltage source $v_{\text{Thermal}}$
in Figure 2.2(b), is

$$v_{\text{Thermal}} = S_{\text{PN}} \Delta T,$$

(2.2)

where $S_{\text{PN}}$ is the effective Seebeck coefficient of the n- and p-type thermoelectric
materials and $\Delta T$ is the temperature difference between the hot and cold ends of the
device.

Bismuth telluride (Bi$_2$Te$_3$), polysilicon (poly-Si), and polysilicon germanium
(poly-SiGe) commonly make up microscale thermal harvesters. Through micromachining
techniques, thin-film- and nanowire-based [88]-[89] Bi$_2$Te$_3$ thermocouples are capable of
converting 15-20 µW/cm$^2$ from 10-20 °C (10-20 K) temperature differences [60], [80].
What is more, a Bi$_2$Te$_3$–based thermal harvester has been co-packaged above a silicon
integrated circuit (IC) die to generate 4 µW/cm² with 60 °C (60 K) gradients [58]-[59]. CMOS-compatible integration into the silicon IC is possible by using poly-Si and poly-SiGe as thermoelectric materials at the expense of lower power generation, which ranges between 1.5-6 and 1-3 µW/cm², respectively, at 5-10 °C (5-10 K) gradients [90]. Although these results are promising, thermal harvesters require temperature differences that, at the tiny scales of a microsystem, are impractical.

Large thermal gradients and proper material selection are essential to maximize voltage and power generation. This is a challenge because significant temperature differences, greater 10 °C (10 K), are unexpected in a microsystem and, as a result, power and voltage levels diminish as differences become smaller with harvester dimensions. Also, the Seebeck coefficients of typical thermoelectric materials (e.g., Bi₂Te₃, poly-Si, and poly-SiGe) are in the order of 100-200 µV/K [87]-[90] that when combined with small temperature gradients result in extremely low voltages per thermocouple that make designing effective and efficient power-conditioning circuits challenging [91]-[93]. Placing several thermocouple elements in a series configuration alleviates the necessity of large thermal gradients (from 500 to 1000 to achieve at least 1 V open circuit voltage with a 10 °C difference), but at the expense of large series electrical resistance, which increases power losses and adversely decreases any gains. The generated voltage can be improved by surrounding the device with a thermally insulating material, such as SiO₂, to maintain heat flow across the thermocouples and avoid heat dissipation through the substrate. Nevertheless, temperature differences greater than 10 °C (10 K) are rare in a microsystem, especially without the aid of additional heat sinks. Therefore, low voltages and power levels are still expected.
2.3 Kinetic Energy

Kinetic energy from motion is available in many forms and environments, including, for instance, cyclic events such as the mechanical vibrations of a motor or engine [94]-[104] and intermittent movements like a person walking [105]-[108]. Ambient vibrations provide a plentiful source of kinetic energy ranging between 1-500 Hz with peak accelerations between 0.1-12 m/s² from diverse environments, such as a car engine (12 m/s² at 200 Hz), the floor of an office building (0.2 m/s² at 100 Hz), and a domestic freezer (0.1 m/s² at 50 Hz) [62], [94]-[95]. The extraction of this energy is based on the movement of a spring-mounted mass relative to its support frame (mass-spring-damper system), as shown in Figure 2.3 [95], [96]. Mechanical acceleration produced by vibrations cause the mass to move and oscillate. This relative displacement causes opposing mechanical damping forces, such as friction, exerted against the mass to absorb kinetic energy and suppress the oscillation. Intentionally imposing an electrical damping force with a magnetic field (electromagnetic), an electric field (electrostatic), or strain on piezoelectric materials harnesses the kinetic energy and converts it to usable electrical energy (with power densities up to 400 µW/cm³) [94]-[95]. The conversion procedure represents a kinetic energy loss, as energy is transformed from mechanical to electrical, and it is therefore regarded as an electrical damping mechanism or, in this case, energy scavenged from the surrounding environment.

A mass-spring-damper system models a vibration-based harvester, where a mass suspended from a spring oscillates as the harvester vibrates (Figure 2.3). Dampers represent the mechanical energy lost, including typical losses like air friction and the energy converted to electricity, and their characteristics depend on the specifics of the
energy transduction mechanism (e.g., electrostatic, electromagnetic, or piezoelectric). Note that energy available from the vibrating source is significantly greater than what the harvester extracts. By assuming that mechanical and electrical damping is linear and viscous (proportional to velocity), the mass of the vibration source is greater than the harvester’s mass, and that it sustains sinusoidal displacement, although not accurate for every case, serves to reach some important fundamental physical relationships. Following these assumptions, the theoretical power extraction $P_{\text{Vibration}}$ as a function of vibration frequency is

$$P_{\text{Vibration}}(\omega) = \frac{\left(\frac{m}{\omega_n}\right)^3 a^2}{\left[1-\left(\frac{\omega}{\omega_n}\right)^2\right]^2 + 2\left(\zeta_E + \zeta_M\right)\left(\frac{\omega}{\omega_n}\right)^2},$$  

(2.3)

where $m$ is the mass, $\omega$ and $a$ are the frequency and acceleration magnitude of the driving vibrations, respectively, $\zeta_E$ and $\zeta_M$ are, correspondingly, the electrically- and mechanically-induced damping ratios, and $\omega_n$ is the natural frequency of the mechanical system [94]-[97]. Increasing $\zeta_E$ and $\zeta_M$ helps spread the bandwidth of the generated energy.
power, which might be better for applications where vibration frequency changes, although at the expense of peak power. The system extracts greatest power when it is at resonance with the fundamental frequency of the ambient vibrations ($\omega_n = \omega$), which is the frequency where maximum acceleration exists. At resonance the extracted power is

$$P_{\text{Generated}}|_{\omega = \omega_n} = \frac{1}{4} \left( \frac{\zeta_E}{\zeta_E + \zeta_M} \right) \left( \frac{m}{\omega_n} \right) a^2,$$

(2.4)

where for a given acceleration, the extracted power is inversely proportional to frequency. Even though the theoretical power generated is derived from simple viscous models and do not consider specifics of the transduction mechanism, which feature nonlinear behavior, it helps to reach general observations. For instance, generated power is proportional to the oscillating mass and square of acceleration (in other words, the environment exerts more force and, thus, does more work) and strongly dependent on resonance to its vibration source.

Irrespective of the transduction means, the kinetic energy harvested from vibrations is constrained by various characteristics that have important consequences in a microsystem implementation. For instance, placing and fabricating a large mass is difficult to realize in microscale spaces, thereby limiting scavenged power. Also, typical sources of vibrations feature peak accelerations below 500 Hz [94]-[95]. Therefore, the kinetic energy harvester must feature a low-frequency oscillating structure that resonates with the source, which is challenging at the diminutive scales of a microsystem. As dimensions decrease, mechanical spring coefficients increase, while large masses become increasingly difficult to integrate, and consequently, the natural frequency of the harvester $\omega_n$ increases.
\[ \omega_n = \sqrt{\frac{k}{m}} \] (2.5)

where \( k \) is the spring coefficient and \( m \) is the mass of the mechanical system. Nevertheless, it is important to consider that in certain applications, with exceedingly strong vibrations, high-frequency harmonics might be sufficient, but not for most practical environments. Finally, parasitic energy extraction (mechanical damping), such as air friction, should be reduced to achieve greater displacement and therefore increase power. A vacuum-packaged oscillating mass produces up to a four-fold increase in scavenged power, but at the expense of added fabrication complexity and cost [110]. In conclusion, harvesting kinetic energy from vibrations can be challenging when integrated into a microscale system, such as a sensing node in a wireless network, but its wide availability and accessibility in many applications and diverse environments merits consideration.

2.4 Radio Frequency (RF) Electromagnetic Energy

Because of today’s widespread use of wireless telecommunications, such as from cellular phones and wireless local area networks (WLAN), it is attractive to scavenge energy from background far-field radio frequency (RF) signals (10 kHz to 10 GHz) [111]-[115]. The problem is ambient RF power and voltage levels are extremely low because propagation energy drops off rapidly as distance from the source increases. To generate any useful power, it is therefore necessary for the harvester to remain close to an RF source. For instance, power densities in the range of 0.01-0.1 µW/cm² are likely when 25-100 m away from a GSM base station (for cellular phones) and a magnitude lower from a WLAN source [61]. Another challenge is that RF ambient energy spreads across a
wide spectrum and requires a large broadband antenna. For this reason, an RF harvester best suits applications where a powerful source exists and it is tuned to the known transmitted frequency. In such cases, for example, 0.22-11.7 µW/cm$^2$ can be extracted from 30-170 µW/cm$^2$ incident RF power at 2.4 GHz [113]. In general, however, it is more likely a nearby RF radiation source is not present, and, consequently, it is difficult to find the benefits of RF energy scavenging over other forms of more available ambient energy sources.

2.5 Comparison

Photovoltaic cells, thermocouples, mechanical vibration-driven microgenerators, and antennae can harness ambient energy from light, temperature gradients, vibrations, and RF radiation. Table 2.2 compares each harvesting technology qualitatively and states their main integration challenges. The power levels that can be achieved, however, depend on the conditions surrounding the application. Light energy, for instance, is exceptionally sensitive to the particular conditions of the system surroundings, with power levels that vary from 15 mW/cm$^2$ outdoors to 10 µW/cm$^2$ for indoor applications. As a result, even though photovoltaic cells can capably generate the highest power density levels, when compared to other harvesting technologies, it is not a dependable source, even if available. Meanwhile, thermal gradients are not sufficiently large to generate practical and usable voltages in a microscale environment without placing numerous cells in series. Mechanical vibrations, however, from engines, flowing water, gusting winds, moving people, ventilation ducts, and so on, are in fact more likely, and consequently, present a more abundant, stable, and predictable energy source in a wide variety of environments and applications.
<table>
<thead>
<tr>
<th>Energy Conversion Method</th>
<th>Advantages</th>
<th>Drawbacks</th>
<th>Challenge in Microsystem</th>
</tr>
</thead>
<tbody>
<tr>
<td>Photovoltaic</td>
<td>- No moving parts, reliable</td>
<td>- Highly dependant on surrounding light conditions</td>
<td>- Small surface areas</td>
</tr>
<tr>
<td></td>
<td>- Mainstream technology</td>
<td></td>
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<tr>
<td></td>
<td>- Scalable</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Thermoelectric</td>
<td>- No moving parts, reliable</td>
<td>- Low conversion efficiency</td>
<td>- Low temperature gradients</td>
</tr>
<tr>
<td></td>
<td>- Scalable</td>
<td>- Low output voltages</td>
<td></td>
</tr>
<tr>
<td></td>
<td>- Durable</td>
<td></td>
<td></td>
</tr>
<tr>
<td>RF Energy</td>
<td>- Possibly many energy sources available</td>
<td>- Low power and voltage levels</td>
<td>- Relatively large broadband antenna</td>
</tr>
<tr>
<td>Vibrations</td>
<td>- Abundant and stable</td>
<td>- Requires a moving mechanical component</td>
<td>- Low resonant frequency</td>
</tr>
<tr>
<td></td>
<td>- Prevalent</td>
<td></td>
<td>- Integration of large mass</td>
</tr>
<tr>
<td></td>
<td>- Predictable (in many applications)</td>
<td></td>
<td></td>
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</tbody>
</table>
CHAPTER 3
VIBRATION-BASED HARVESTERS

In many environments and applications, vibrations offer a plentiful and consistent source of kinetic energy. Although vibrations may not generate as much power as sunlight, they consistently and reliably produce considerably more than indoor lighting and thermal gradients. Intentionally imposing an electrical damping force with a magnetic field (electromagnetic), an electric field (electrostatic), or strain on a piezoelectric material harnesses the kinetic energy from mechanical vibrations and converts it into electrical energy. Each technique, however, offers different benefits as well as technical challenges that determine its viability and potential for integration into self-powered microsystems.

3.1 Electromagnetic Harvesters

Electromagnetic energy harvesters use a magnetic field to convert mechanical energy to electrical [110], [116]-[137]. By exploiting Faraday’s law, a coil generates a voltage as the flux of a magnetic field changes with time. For instance, a coil can be attached to an oscillating mass that moves in response to vibrations through a magnetic field established by a stationary and permanent magnet, as seen in Figure 3.1(a). As the coil moves, it travels through a varying amount of magnetic flux, inducing an ac voltage on the coil. Conversely, the same effect will result if a magnetic mass resonates with vibrations as the coil remains fixed, which in many instances is preferred since it is easier to connect the stationary coil to other system electronics and to utilize the magnet as the seismic mass. The generated open-circuit voltage, modeled as \( v_{EM} \) with series resistance
Rs in Figure 3.1(b), will depend on the number of coil turns N, length of each coil l, strength of the magnetic field B, and total relative displacement y of the coil movement

$$v_{EM}(t) = NBI\frac{\partial y}{\partial t}.$$ (3.1)

Given the small amplitude of the vibrations, the induced ac voltage is inherently small, in the order of 100-300 mV [118], and must therefore be increased to be a feasible solution compatible with the threshold voltages of common electronic circuits (about 700 mV). Methods to increase the induced voltage include using a transformer, increasing the number of turns of the coil, and/or increasing the permanent magnetic field strength. However, all options are limited by the size constraints of the intended microscale application. What is more, the ac voltage generated requires rectification and proper power conditioning, both of which impose losses in the system, and is particularly challenging due to the extremely low voltages produced. Furthermore, electromagnetic harvesters require integration of magnetic materials, such as neodymium iron boron (NdFeB), which are difficult to miniaturize.

Figure 3.1 (a) Electromagnetic vibration energy harvester and (b) its circuit model.
3.2 Piezoelectric Harvesters

Piezoelectric harvesters convert mechanical energy to electrical by straining a piezoelectric material [95], [138]-[178]. A piezoelectric material features aligned dipoles within its crystal structure so that if a voltage is applied to the material in the same, or opposite, direction as the polarity of the dipoles, the material elongates, or shrinks, in that direction. Conversely, if applied stress on the material produces strain in the direction of the dipoles polarity, a charge separation occurs across the material, and hence a voltage is produced. In other words, strain, or deformation, in a piezoelectric material causes charge separation across the device, producing an electric field and, consequently, voltage drop $v_{PZ}$ proportional to the stress applied

$$v_{PZ}(t) = \sigma(t) \frac{t_h d_z}{\varepsilon},$$

(3.2)

where $d_z$ is the piezoelectric strain coefficient, $\sigma$ is the applied mechanical stress, $\varepsilon$ is the dielectric constant, and $t_h$ is the thickness of the material.

The oscillating system is typically a cantilever beam structure since it provides higher strain and lower resonance for a given input force [159]. The beam, shown in Figure 3.2(a), is constructed as a two-layer bimorph structure with two piezoelectric sheets separated by a center dielectric shim. The cantilever is fixed on one end to a stationary base while a moving mass is attached on the other. As the mass resonates, one side of the cantilever is in tension while the other is in compression, constantly expanding and contracting, creating a changing ac voltage within each layer. If the piezoelectric material in each layer is poled in opposite directions, the generated voltages add and are coupled through the center dielectric shim. The cantilever design also permits harnessing energy from a wider vibration frequency spectrum by fabricating several cantilevers with
different resonant frequencies in the same substrate [149]-[153]. Another possibility is to combine the piezoelectric harvester with an electromagnetic one by including a magnetic mass at the tip of the cantilever [154].

Piezoelectric materials are typically deposited onto the substrate as thin films that are compatible with microfabrication processes [155]-[156]. However, thin-film piezoelectric layers suffer from greatly reduced mechanical-to-electrical coupling efficiency (which directly relates to the piezoelectric strain coefficient of the material). Lead zirconate titanate (PZT) is the most common piezoelectric material used because of its relatively higher electro-mechanical coupling efficiency (up to 75%) [62], [157]-[161]. PZT, however, is brittle and can only support small masses. Other materials such as polyvinylidene flouride (PVDF) and aluminum nitride (AlN) offer a more robust and flexible solution, although at the expense of significantly lower coupling (15%) [162]-[163]. Lead magnesium niobate-lead titanate (PMN-PT) has even higher coupling efficiencies (92%), but is more expensive [156], [164].

Figure 3.2. (a) Piezoelectric bimorph beam and (b) its circuit model.
The piezoelectric cantilever generated voltage $v_{\text{Piezo}}$ will vary with time and strain, effectively producing an irregular ac signal, as the circuit model in Figure 3.2(b) shows. Its magnitude depends on the mechanical stress applied and the piezoelectric strain coefficient of the material, and is in the order of several volts (3-10 V) [94]-[95], a magnitude greater than the voltages produced by electromagnetic harvesters. A shortcoming of piezoelectric converters, however, is the additional circuitry required to rectify and convert the extracted power from an unsteady, high impedance source to a stable, low impedance supply, which incurs additional power losses and therefore decreases the overall effectiveness and efficiency of the conversion mechanism [165]-[178]. In summary, piezoelectric energy conversion produces relatively better voltage levels with higher power density than the electromagnetic approach, but retain the requirements of rectification, power conditioning, and integration of an exotic material.

3.3 Electrostatic Harvesters

An electrostatic harvester harnesses energy from the work done by vibrations against the electrostatic force of a vibration-dependent variable capacitor $C_{\text{VAR}}$ (i.e., a varactor) [95], [179]-[199]. Mechanical energy is converted to electrical energy as the capacitance of the charged capacitor decreases in response to externally applied vibrations. From a physical standpoint, vibrations cause the gap distance and/or overlap area of a parallel-plate capacitor to vary with a net effect, under constant charge or voltage conditions, of producing electrical energy.

3.3.1 Charge-Constrained Harvesters

One way of harvesting energy is by constraining charge $Q_{C}$ in $C_{\text{VAR}}$ by leaving it open circuited, so that as vibrations separate its plates, capacitance decreases and
capacitor voltage $v_C$ increases, augmenting the energy stored in $C_{VAR}$ in the process with energy from the environment

$$v_C = \frac{Q_C}{C_{VAR}}.$$  \hspace{1cm} (3.3)

Since charge stays constant, the capacitance variation amplifies $C_{VAR}$’s initial voltage by its maximum ($C_{MAX}$) to minimum ($C_{MIN}$) ratio to a final value of

$$V_{C,\text{Final}} = \left(\frac{C_{MAX}}{C_{MIN}}\right)V_{C,\text{Initial}}.$$  \hspace{1cm} (3.4)

In other words, the increasing squaring effects of voltage $v_C$ on the energy $E_C$ stored by $C_{VAR}$ offset the decreasing linear effects of capacitance,

$$E_C = \frac{1}{2} C_{VAR} v_C^2.$$  \hspace{1cm} (3.5)

Therefore, the net electrical energy gain stored in $C_{VAR}$, or converted mechanical energy, through a single maximum-to-minimum cycle ($\Delta C_{VAR}$) is

$$\Delta E_{\text{Gain}} = E_{C,\text{Final}} - E_{C,\text{Initial}} = \frac{1}{2} \left(\frac{C_{MAX}}{C_{MIN}}\right) \Delta C_{VAR} V_{C,\text{Initial}}^2,$$  \hspace{1cm} (3.6)

which is the difference between the final and initial energy stored in the capacitor. Note that one drawback is that $C_{VAR}$ must be initially charged with energy $E_{C,\text{Initial}}$.

The maximum voltages charge-constrained systems produce, however, surpass the breakdown limits of most modern CMOS or BiCMOS technologies by a considerable margin. For example, a 1-200 pF variation amplifies the initial voltage across $C_{VAR}$ by a factor of 200 producing peak voltages of roughly 25-200 V from inputs as low as 0.125-1 V. Even greater voltages can be expected if $C_{VAR}$ is charged with an initial voltage derived from a Li Ion (2.7-4.2 V). More expensive and specialized technologies, such as silicon-on-insulator (SOI) CMOS processes [183]-[184], can sustain these voltage
extremes but their increased costs limit the extent to which the market will adopt them, especially in wireless microsensors where volume production and low cost are driving factors. If instead, the initial voltage of the capacitor is kept sufficiently low, so that as capacitance decreases its voltage increases up to the levels of the in-package battery $V_{\text{BAT}}$ (2.7-4.2 V), the energy gain is

$$\Delta E_{\text{Gain}} = \frac{1}{2} \left( \frac{C_{\text{MIN}}}{C_{\text{MAX}}} \right) \Delta C_{\text{VAR}} V_{\text{BAT}}^2. \quad (3.7)$$

In this manner, constraining charge is compatible with low-cost, high-volume processes, but at the expense of significantly reducing the theoretical energy gain (maximum-minimum capacitance ratio is inverted). Still, additional circuitry is required to transfer energy to and from the capacitor between differing voltage levels. Holding voltage constant, however, provides a more viable solution that is compatible and benign to mainstream microchip technologies.

### 3.3.2 Voltage-Constrained Harvesters

By restraining voltage, the mechanical energy moving the capacitor plates and decreasing capacitance drives charge out of the capacitor, generating harvesting current $i_{\text{HARV}}$

$$i_{\text{HARV}} = \frac{dq}{dt} = C \frac{\partial V}{\partial t} + V \frac{\partial C}{\partial t} = V \frac{\partial C}{\partial t}. \quad (3.8)$$

Although constraining voltage is compatible with standard processes, typical implementations employ an additional voltage source (e.g., capacitor, electret, etc.) to fix $C_{\text{VAR}}$’s voltage [179], which contradicts the goals of integration, and an energy-transferring circuit to charge the battery. Connecting $C_{\text{VAR}}$, after being charged by $V_{\text{BAT}}$, to a constraining capacitor (i.e., a low-capacity battery) [190]-[192], via a unidirectional
diode holds $C_{VAR}$’s voltage, but only momentarily because $i_{HARV}$ raises the constraining capacitor’s voltage, so $C_{VAR}$ must undergo a charge-constrained phase every cycle to keep up. Although viable, the basic problem here is that energy harvested during that charge-constrained phase does not charge the targeted battery, leaving otherwise useful energy unharvested in the variable capacitor, as only a fraction of the full capacitance variation is now harvested.

Another possibility is to embed a material that has a permanent charge separation and, thus, a constant voltage, as in the case of electrets and charged electrodes [193]-[199]. A permanent-charge device, however, represents a voltage reference and is not a true low-impedance voltage source capable of delivering current or storing energy since its charges are trapped. By inducing a constant voltage across the capacitor, however, $C_{VAR}$ generates an ac harvesting current, which flows into and from the load as $C_{VAR}$ increases and decreases. Disconnecting the permanent-charge material when $C_{VAR}$ increases can help channel only positive harvesting current to the load, but when reconnected it cannot recharge $C_{VAR}$ to re-establish its electrostatic force (since charges are fixed). Additionally, they require a complicated assembly process to integrate two separate substrates, one being the electret, or additional fabrication sequences to charge the material via electron tunneling that can result in higher costs. Nevertheless, trapped-charge devices still embody an additional voltage source. Constraining $C_{VAR}$’s voltage with the already-existing system battery (i.e., thin-film Li Ion) that is to be ultimately charged, on the other hand, enhances integration because no additional source is required.
3.3.3 Variable Capacitor

The most attractive feature of electrostatic energy harvesting is its IC-compatible nature. Variable capacitors based on microelectromechanical systems (MEMS) technology are routinely fabricated for sensing and actuation applications through relatively mature silicon micromachining techniques, such as deep reactive ion etching (DRIE) [95], [200]. The movement of the capacitor plate is preferably in-plane (in parallel) with the substrate to avoid problems associated with out-of-plane capacitors, seen in Figure 3.3(a), such as large mechanical damping and surface interactions. The capacitor plates for in-plane motions are typically fabricated with interdigitated fingers in a comb structure, where varying overlap area, shown in Figure 3.3(b), provides greater plate displacements and can achieve low minimum capacitances (~0.1 pF – parasitic-limited), but suffers from stability problems. Changing gap distance, illustrated in Figure 3.3(c), alternatively, provides better stability due to smaller spring deflections and large maximum capacitances (~800 pF) but at the expense of greater minimum capacitances (~20 pF) [95]. Fabricated prototypes in literature typically include capacitances ranging between 400 and 50 pF [97], [179], [181]-[182], [188], with a maximum ΔC\text{VAR} of 1570

Figure 3.3. (a)-(c) Possible physical implementations of harvesting variable capacitor C\text{VAR} and (d) its circuit model.
to 62 pF [189]. Regardless of the mechanical design of the variable capacitor, its plates will generate parasitic series resistance, while parasitic capacitances, which require more initial charge, will develop between the moving plate and the substrate and sidewalls, as the electrical model in Figure 3.3(d) illustrates.

### 3.4 Comparison

Energy from ambient mechanical vibrations can be harvested by harnessing the work done against the damping forces produced by magnetic fields, electric fields, and strain on piezoelectric materials. A fair comparison between published state-of-the-art vibration harvesters in terms of generated power density levels is difficult because of the wide variety of sizes, materials, resonant frequencies, and vibration conditions (e.g., acceleration) that are published and evaluated. Furthermore, some results do not consider the required power conditioning electronics (such as low-voltage rectifiers for electromagnetic harvesters), only account part of the volume occupied by the device (by not including the mass of a piezoelectric harvester), or do not count the initial energy investment required to establish $C_{VAR}$’s electrostatic force. Nevertheless, several researchers have attempted to normalize and compare technologies [62], [97]-[98], where, in general, at small volumes (less than 1 cm$^3$) electromagnetic harvesters feature power densities around 50-100 µW/cm$^3$, although a 310 µW/cm$^3$ harvester has been published [118], while electrostatic harvesters feature similar densities in the 10-100 µW/cm$^3$ range [179], [192]. Piezoelectric harvesters, however, provide the greatest power densities, up to 400 µW/cm$^3$ [159]. Commercial harvesting products mostly focus on electromagnetic systems (e.g., Perpetuum, Ferro Solutions), although some piezoelectric products exist (e.g., Mide) [98]. These are bulky (40-200 cm$^3$) and expensive, and,
therefore, target niche applications where size and cost are not as important as long lifetimes (e.g., military applications). Since energy harvesting is still a young and immature technology, it is better to evaluate each technique from a more qualitative and practical point of view.

Each vibration-based harvesting technique offers different advantages and drawbacks, which Table 3.1 summarizes. By comparing each in terms of integration potential, electromagnetic and piezoelectric scavengers are less attractive because both methods require the incorporation of difficult-to-integrate magnetic and piezoelectric materials. Electrostatic harvesters, in contrast, are fully compatible with mature MEMS technologies and capable of generating moderate power levels without the use of exotic materials or obscure process steps. Even then, they require an initial charge to exist in the capacitor as an investment to initiate the harvesting process, which must be a fraction of the net energy gain. Additionally, the harvester must synchronize with the vibration source.

Nevertheless, out of the electrostatic energy harvesting techniques (voltage- or charge-constrained), holding voltage constant is more compatible with mainstream technologies because, unlike in charge-constrained schemes, voltage-constraining harvesting capacitor $C_{VAR}$ protects the circuit from voltages that exceed the breakdown limits of standard CMOS and BiCMOS technologies. As a result, voltage-constrained electrostatic harvesters present a viably cheaper alternative that permits the high-volume fabrication necessary, for example, to deploy numerous sensing nodes in a wireless network. Some method is still required to hold voltage constant without adding another source or utilizing unusual materials. It is for these reasons that the system to be
discussed in subsequent chapters harnesses kinetic energy with a voltage-constrained electrostatic harvester, in which the power cache (Li Ion) that stores the scavenged energy and caters the power-demanding functions of the microsystem also holds the $C_{VAR}$’s voltage constant. In this way, the energy-harvesting integrated circuits to be presented avoid an additional source by using the system’s already-existing battery and driving $i_{HARV}$ directly into it without intervening microelectronics.
CHAPTER 4
BATTERY-CONSTRAINED HARVESTER

Although the environment offers many sources of energy, kinetic energy from vibrations is abundant, consistent, and in many cases, predictable and systematic. An electrostatic harvester can potentially extract this energy with mainstream MEMS variable capacitors, without the need of atypical materials as in electromagnetic and piezoelectric harvesters. Constraining the variable capacitor’s voltage, rather than charge, eliminates the burden of high voltage, notwithstanding that additional synchronizing circuitry is still necessary. However, by using the system’s battery to hold the capacitor’s voltage constant allows compatibility with standard low-voltage CMOS or BiCMOS semiconductor technologies, circumvents the need for additional voltage sources, and drives the harvested energy directly to the intended energy storage device.

4.1 Physical Analysis

Ambient vibrations represent an abundant source of kinetic energy, which an electrostatic energy scavenger extracts by holding the charge or voltage of a vibration-dependent variable capacitor constant. Figure 4.1 illustrates parallel-plate variable capacitor $C_{VAR}$, where one of its plates moves in response to vibrations, and, in this case, gap distance $z$ varies accordingly. As vibrations cause the plates of charged $C_{VAR}$ to separate, they work against the electrostatic force $F_E$ of the capacitor and mechanical energy converts, or transfers, to the electrical domain.
Fundamentally, $C_{\text{VAR}}$’s moving plate represents the oscillating mass from the mass-spring system presented in Figure 2.3. As a result, vibrations must also overcome and work against other forces present in the system, such as air friction and the spring potential force shown in Figure 4.2, that do not convert energy to electricity. For instance, the exchange between potential energy in the spring and kinetic energy causes the capacitor’s plate (i.e., seismic mass) to oscillate at a resonant frequency, much like the energy exchange between an inductor and a capacitor in a resonant LC circuit. Accordingly, forces such as friction dissipate energy from vibrations and attempt to dampen oscillations, behaving like a resistor. Harvesting energy, however, represents a diminutive fraction of the energy available from typical vibrations, and do not significantly diminish the oscillations of the parallel-plate variable capacitor.

Figure 4.1. Vibration-driven variable capacitor $C_{\text{VAR}}$.

Figure 4.2. Harvesting energy from $V_{\text{BAT}}$-constrained variable capacitor $C_{\text{VAR}}$. 
When leaving $C_{VAR}$ open circuited to hold capacitor charge $Q_C$ constant, its electrostatic force $F_E$, which is independent of $z$, is

$$F_E(z) = \frac{Q_C^2}{2\varepsilon_0 A}, \quad (4.1)$$

where $A$ represents the area of the plates, and $\varepsilon_0$ is the dielectric permittivity of air ($\varepsilon_0 \approx 8.85 \text{ pF/m}$). The energy converted from the environment ($E_{CONV}$) is the work done by the vibrations to separate the capacitor plates from $z_{MIN}$ to $z_{MAX}$ (or in other words, capacitance decreases from $C_{MAX}$ to $C_{MIN}$), which is

$$E_{CONV} = \text{Work} = \int F_E(z) dz = \frac{Q_C^2}{2\varepsilon_0 A} \int_{z_{MIN}}^{z_{MAX}} dz = \frac{Q_C^2}{2} \left( \frac{1}{C_{MIN}} - \frac{1}{C_{MAX}} \right) \quad (4.2)$$

and simplifies to

$$E_{CONV} = \frac{1}{2} (V_{BAT} C_{MAX})^2 \left( \frac{C_{MAX} - C_{MIN}}{C_{MAX} C_{MIN}} \right) = \frac{1}{2} \Delta C_{VAR} V_{BAT}^2 \left( \frac{C_{MAX}}{C_{MIN}} \right), \quad (4.3)$$

where $C_{VAR}$ is initially charged to battery voltage $V_{BAT}$ before vibrations separate the plates. The converted mechanical energy is stored in $C_{VAR}$, as confirmed in Equation 3.6, which shows that capacitor energy augments by $E_{CONV}$.

Conversely, when restraining capacitor voltage $v_C$ to $V_{BAT}$, $F_E$ as a function of $z$ is

$$F_E(z) = \frac{q_C^2}{2\varepsilon_0 A} = \left( \frac{V_{BAT} C}{2\varepsilon_0 A} \right)^2 = \frac{V_{BAT}^2}{2\varepsilon_0 A} \left( \frac{\varepsilon_0 A}{z} \right)^2 = \frac{\varepsilon_0 A V_{BAT}^2}{2z^2}. \quad (4.4)$$

Therefore, the converted energy, which equals the work done by the vibrations, is

$$E_{CONV} = \int F_E(z) dz = \frac{\varepsilon_0 A V_{BAT}^2}{2} \int_{z_{MIN}}^{z_{MAX}} \frac{1}{z^2} dz = \frac{\varepsilon_0 A V_{BAT}^2}{2} \left( \frac{1}{z_{MIN}} - \frac{1}{z_{MAX}} \right), \quad (4.5)$$

and simplifies to

$$E_{CONV} = \frac{1}{2} \Delta C_{VAR} V_{BAT}^2, \quad (4.6)$$
which, for the same $C_{\text{MAX}}$ to $C_{\text{MIN}}$ variation ($\Delta C_{\text{VAR}}$), is less than what is converted when constraining charge.

Charge-constrained harvesters, though, allow $v_C$ to increase to significant levels incompatible with low-voltage semiconductor processes. Alternatively, charging $C_{\text{VAR}}$ to a low voltage, so that $v_C$ increases to $V_{\text{BAT}}$, results in the converted energy reducing to

$$E_{\text{CONV}} = \frac{1}{2} (V_{\text{BAT}} C_{\text{MIN}})^2 \left( \frac{C_{\text{MAX}} - C_{\text{MIN}}}{C_{\text{MAX}} C_{\text{MIN}}} \right) = \frac{1}{2} \Delta C_{\text{VAR}} V_{\text{BAT}}^2 \left( \frac{C_{\text{MIN}}}{C_{\text{MAX}}} \right). \quad (4.7)$$

For the same capacitance variation, however, less energy is converted than in the voltage-constrained case, and, as a result, for low-voltage CMOS or BiCMOS compatibility, voltage-constrained harvesters offer the best solution. Accordingly, a trade-off between energy gain and cost exists, where the greater energy conversion of charge-constrained systems comes at the expense of using pricier processes that are tolerant to high voltages. Yet, regardless of the method, electrostatic harvesters require that $C_{\text{VAR}}$ is initially charged at $C_{\text{MAX}}$ to establish the electrostatic force that vibrations will work against.

### 4.2 Energy-Harvesting Scheme

The presented voltage-constrained energy-harvesting system features a rechargeable battery (thin-film Li Ion) that both clamps the voltage of a vibration-driven variable capacitor and stores the harvested energy. The process operates in separate steps, as illustrated in Figure 4.3, where, first, the battery invests an initial amount of energy to precharge capacitor $C_{\text{VAR}}$ to the battery voltage $V_{\text{BAT}}$ when its capacitance is highest ($C_{\text{MAX}}$). This investment constitutes an energy loss in the system (from the battery), which assuming no other losses exist in the transfer, is

$$E_{\text{INV}} \equiv E_C = \frac{1}{2} C_{\text{MAX}} V_{\text{BAT}}^2. \quad (4.8)$$
Mechanical forces from ambient vibrations then work against the capacitor's established electrostatic force and cause $C_{VAR}$ to decrease, converting mechanical energy to electrical in the process. As the capacitance of $C_{VAR}$ decreases, the battery holds constant the capacitor voltage $v_C$ at $V_{BAT}$, generating harvesting current $i_{HARV}$, shown in Equation 3.8, which charges the battery to gain energy $E_{HARV}$

$$E_{HARV} = \int V_{BAT}i_{HARV}(t)dt = V_{BAT}^2\int \frac{dC(t)}{dt}dt = \Delta C_{VAR}V_{BAT}^2.$$ (4.9)

After reaching minimum capacitance $C_{MIN}$, $C_{VAR}$ disconnects from the battery to avoid the reverse process where current discharges the battery. Instead, the remnant energy in $C_{VAR}$ ideally transfers back to the battery, effectively recovering energy $E_{REC}$

$$E_{REC} = \frac{1}{2}C_{MIN}V_{BAT}^2.$$ (4.10)
Subsequently, the capacitor, exhausted of charge and energy, resets to its original state of maximum capacitance without the system incurring any energy loss or gain. Note that although \( C_{\text{VAR}} \) lacks an electrostatic force, vibrations still work against other forces, such as friction, to push its plates back towards each other. Summing all the gains to the initial investment yields a net energy profit, equal to the converted mechanical energy from the environment (Equation 4.6), of

\[
E_{\text{NET}} = E_{\text{HARV}} + E_{\text{REC}} - E_{\text{INV}} = \frac{1}{2} \Delta C_{\text{VAR}} V_{\text{BAT}}^2. \tag{4.11}
\]

During the harvesting phase, the battery gains energy \( E_{\text{HARV}} \), which is composed of the converted mechanical energy \( E_{\text{CONV}} \) (from Equation 4.6) plus part of the invested electrical energy stored in the capacitor. While capacitance decreases, the capacitor’s charge, which was established during the precharge phase, decreases by

\[
\Delta q_C = \int i_{\text{HARV}}(t) \, dt = \Delta C_{\text{VAR}} V_{\text{BAT}} \quad \tag{4.12}
\]

and, consequently, the energy in \( C_{\text{VAR}} \) reduces by

\[
\Delta E_C = \frac{1}{2} \Delta q_C V_{\text{BAT}} = \frac{1}{2} \Delta C_{\text{VAR}} V_{\text{BAT}}^2. \quad \tag{4.13}
\]

Accordingly, the harvested energy \( E_{\text{HARV}} \) is twice the energy given up by \( C_{\text{VAR}} \), where the difference represents the mechanical energy input to the system. In effect, part of the charge transferred during precharge to \( C_{\text{VAR}} \) returns to \( V_{\text{BAT}} \), and therefore, \( V_{\text{BAT}} \) earns this energy back, along with converted mechanical energy \( E_{\text{CONV}} \)

\[
E_{\text{HARV}} = \Delta E_C + E_{\text{CONV}} = \Delta C_{\text{VAR}} V_{\text{BAT}}^2. \tag{4.14}
\]

The remaining invested energy (from remaining charge in \( C_{\text{VAR}} \)) returns to \( V_{\text{BAT}} \) during the recovery phase.

The process of transferring energy, however, is not lossless and the system therefore loses energy in all steps, which is one of the fundamental challenges of
harvesters. When $C_{\text{VAR}}$ reaches $C_{\text{MIN}}$, the energy recovered ($E_{\text{REC}}$) is a small fraction of the total energy harvested and on the same order in magnitude, if not less, as the energy required to salvage it, defeating the intent of a recovery phase. To this end, it is better to omit recovery, leaving the capacitor open-circuited in a charge-constrained state during reset. Consequently, as vibrations force $C_{\text{VAR}}$ to increase again, its voltage naturally resets and decreases to a substantially lower voltage near zero, that is, to

$$V_{C_{\text{MIN}}} = \frac{C_{\text{MIN}}}{C_{\text{MAX}}} V_{\text{BAT}}.$$  \hspace{1cm} (4.15)

Nevertheless, the energy harvested still exceeds the investment, leaving a theoretical net gain of

$$E_{\text{NET}} = E_{\text{HARV}} - E_{\text{INV}} = \left(\frac{1}{2} C_{\text{MAX}} - C_{\text{MIN}}\right) V_{\text{BAT}}^2.$$  \hspace{1cm} (4.16)

When $C_{\text{VAR}}$ reaches $C_{\text{MAX}}$, the harvester again precharges $C_{\text{VAR}}$ and the process repeats, as Figure 4.4 shows.

Figure 4.4. Battery-constrained energy-harvesting scheme without recovery and a charge-constrained reset phase.
Omitting recovery forfeits recuperation of the remnant energy on $C_{\text{VAR}}$ when it reaches $C_{\text{MIN}}$. Vibrations now work along the remaining electrostatic force and, in essence, convert the remnant electrical energy back to the physical domain. In other words, energy returns to the environment. However, the harvesting system still uses this energy, as the decreasing capacitor voltage indicates when $C_{\text{MAX}}$ is reached (at $V_{C,\text{MIN}}$). The control circuit therefore uses the unrecovered remnant energy to determine when the subsequent precharge phase commences, diminishing its negative impact on the system’s net energy gain per cycle. Not taking advantage of this effect increases the complexity and power requirements of the control circuit by otherwise requiring a capacitance-sensing block. Yet, a recovery phase could be viable if $V_{\text{BAT}}$ gains sufficient remnant energy, after the losses incurred by transferring it, to justify the additional circuit complexity.

4.3 Summary

Electrostatic energy harvesters impose an electrostatic force in vibration-sensitive variable capacitor $C_{\text{VAR}}$, which vibrations work against to convert kinetic energy to electrical. Specifically, holding $C_{\text{VAR}}$’s voltage generates harvesting current $i_{\text{HARV}}$, which in the proposed scheme directly charges the microsystem’s battery that ultimately stores the harnessed energy and delivers it to other system loads when needed. However, an investment $E_{\text{INV}}$ from the battery is required during each vibration cycle to establish $C_{\text{VAR}}$’s electrostatic force. Consequently, as long as the energy harvested $E_{\text{HARV}}$ exceeds the investment, the battery gains energy during each cycle.

In practice, $E_{\text{HARV}}$ must overcome not only precharge investment $E_{\text{INV}}$ but also the losses associated with each phase in the cycle. This is a considerable challenge because
the energy harvested every vibration cycle is low to begin with and each phase requires monitoring and control circuitry to detect $C_{\text{MAX}}$ during reset to initiate the precharge phase and $C_{\text{MIN}}$ to disconnect $C_{\text{VAR}}$ from the battery at the end of the harvesting phase. In addition, precharging $C_{\text{VAR}}$ requires greater energy from $V_{\text{BAT}}$ than $E_{\text{INV}}$ to surmount losses in the transfer. To mitigate these losses, the presented system employs a quasi-lossless inductor-based precharger to transfer $E_{\text{INV}}$ from $V_{\text{BAT}}$ efficiently and preserve the battery’s net energy gain.
CHAPTER 5
HIGH ENERGY GAIN STRATEGIES

One of the fundamental challenges of harvesters is that to create a net energy gain and increase the battery’s state of charge, energy dissipated by the system should not exceed the energy harvested. Electrostatic harvesters in particular require a charged variable capacitor, so that vibrations work against the established electrostatic force and convert ambient mechanical energy to electrical. Charging \( C_{\text{VAR}} \) when at maximum capacitance \( C_{\text{MAX}} \) therefore requires an investment from the system’s battery that must be transferred with high efficiency to preserve a net energy gain. Specifically, an inductor-based circuit precharges \( C_{\text{VAR}} \) to \( V_{\text{BAT}} \) without incurring excessive losses that cancel out any gains the system makes. In addition, generated harvesting current \( i_{\text{HARV}} \) channels directly to \( V_{\text{BAT}} \) through a transistor-based switch that avoids the losses a diode might incur.

5.1 Precharger Circuit

When considering energy-transfer strategies to precharge \( C_{\text{VAR}} \) to \( V_{\text{BAT}} \), the most obvious and straightforward embodiment is to connect both with a switch, as Figure 5.1 shows. However, because of substantial concurrent voltages and currents across the switch, this approach is fundamentally lossy and ineffective. For instance, charging \( C_{\text{VAR}} \) from zero by closing switch \( S_{\text{PCH}} \), induces current \( i_{\text{BAT}} \)

\[
i_{\text{BAT}}(t) = \frac{V_{\text{BAT}} - V_{\text{Initial}}}{R_{\text{SW}}} e^{-\frac{t}{R_{\text{SW}}C_{\text{MAX}}}}
\]

(5.1)

To increase \( V_{\text{C}} \)
\[ v_c(t) = V_{BAT} - (V_{BAT} - v_{initial}) \cdot e^{-\frac{t}{R_{SW}C_{MAX}}}. \] (5.2)

As a result, irrespective of its resistance \( R_{SW} \), \( S_{PCH} \) dissipates energy \( E_{SW} \)

\[ E_{SW} = \int_{0}^{\infty} i_{BAT}(t)(V_{BAT} - v_c(t)) dt = \frac{1}{2} C_{MAX} (V_{BAT} - v_{initial})^2 \] (5.3)

that is equal to the initial investment needed in \( C_{VAR} \) if initially discharged. In other words, the battery must invest twice \( E_{INV} \) (Equation 4.8) to charge \( C_{VAR} \) to \( V_{BAT} \) from zero – that is, 50% efficiency. No energy is lost in theory, though, when channeling \( E_{INV} \) to \( C_{VAR} \) via a transfer inductor.

### 5.1.1 Four-Switch Inductor-Based Precharger

Since inductors are ideally lossless energy-storage devices, to maximize energy gain, the battery charges \( C_{VAR} \) to \( V_{BAT} \) with an efficient inductor-based precharger, such as the four-switch circuit seen in Figure 5.2(a). The precharge phase, which occurs at \( C_{MAX} \) before the onset of the harvesting phase, is therefore decomposed into a sequence of two steps: \( V_{BAT} \) first energizes inductor \( L \) with \( E_{INV} \), which then de-energizes to \( C_{VAR} \), as Figure 5.2(b) illustrates. Inductor \( L \) is energized by superimposing battery voltage \( V_{BAT} \) across it with switches \( S_{E1} \) and \( S_{E2} \). Inductor current \( i_L \) consequently increases linearly until sufficient energy is stored, at which point \( S_{E1} \) and \( S_{E2} \) open. After a dead
time, where all switches are turned off to prevent short-circuit conditions, $S_{D1}$ and $S_{D2}$ close and deliver the stored energy to $C_{VAR}$. Once the capacitor voltage reaches that of the battery, $S_{D1}$ and $S_{D2}$ open and the precharge process is complete, whereupon $C_{VAR}$ is ready for the beginning of the harvesting phase.

To ensure enough energy ($E_{INV}$) is transferred from $V_{BAT}$ to $C_{VAR}$, $L$ is energized for a set length of time $t_E$, during which, inductor current $i_L$ increases linearly to a maximum value of

$$I_{L,PK} = \frac{V_{BAT}}{L} t_E.$$  \hspace{1cm} (5.4)

The amount of energy transferred to inductor $L$ is thus

$$E_L = \frac{1}{2} L \cdot I^2_{L,PK} = \frac{1}{2L} (V_{BAT} t_E)^2$$ \hspace{1cm} (5.5)

and equating this to the invested energy required to precharge $C_{VAR}$ ($E_{INV}$), as derived in Equation 4.8, yields a precharge time of

$$E_L = E_{INV} \rightarrow t_E = \sqrt{\frac{L}{C_{MAX}}}.$$

which, assuming a lossless transfer, is independent of $V_{BAT}$. Consequently, even as $V_{BAT}$ changes (a Li Ion typically spans 2.7-4.2 V across its state of charge), a constant $t_E$ transfers sufficient energy to $C_{VAR}$. Another key feature is that remnant energy at $C_{MIN}$

---

Figure 5.2. (a) Four-switch inductor-based precharger and (b) its connections while energizing and de-energizing $L$. 

---
can be recovered, if desired, by using the same inductor circuit, but with an inverted switching sequence. In practice, however, precharge is not free of losses and for that reason \( t_E \) is set slightly higher to offset the energy losses associated with the transfer.

Once \( L \) has sufficient energy, it de-energizes to \( C_{\text{VAR}} \), where, assuming \( C_{\text{VAR}} \) is originally discharged to zero and as derived in Appendix A, capacitor voltage \( v_C \) and current \( i_C \) are

\[
v_C(t) = I_{L,PK} \sqrt{\frac{L}{C_{\text{MAX}}}} \sin(\omega_{LC}t) \tag{5.7}
\]

and

\[
i_C(t) = I_{L,PK} \cos(\omega_{LC}t), \tag{5.8}
\]

where \( \omega_{LC} \) is the LC’s natural resonant frequency, which is

\[
\omega_{LC} = \frac{1}{\sqrt{LC_{\text{MAX}}}}. \tag{5.9}
\]

During this step, the precharger is essentially an LC resonant circuit, where \( L \) and \( C_{\text{VAR}} \) exchange energy between each other. Disconnecting \( L \) when fully de-energized (\( i_L \) is zero) corresponds to \( v_C \) at \( V_{\text{BAT}} \), which occurs after time \( t_D \)

\[
t_D = \frac{\pi}{2} \sqrt{\frac{LC_{\text{MAX}}}{}}. \tag{5.10}
\]

Accordingly, the total precharge time \( t_{\text{PCH}} \), ideally, is

\[
t_{\text{PCH}} = t_E + t_D = \left(1 + \frac{\pi}{2}\right) \sqrt{\frac{LC_{\text{MAX}}}{}} \tag{5.11}
\]

In practice, de-energizing time is longer, since more energy is invested to overcome losses. Precharge ends when \( v_C \) reaches \( V_{\text{BAT}} \), when all switches turn off, and excess energy in the inductor returns to the battery when connected to \( C_{\text{VAR}} \) during harvesting
phase. It is best to minimize this extra energy to reduce the losses associated with transferring energy through the system.

5.1.2 Two-Switch Inductor-Based Precharger

The precharger shown in Figure 5.2(a) features four switches, each of which dissipates energy in any practical implementation. A more efficient solution is to simplify the precharger to two switches, as shown in Figure 5.3(a). The battery energizes both L and \( C_{VAR} \) with \( E_{INV} \) when closing energizing switch \( S_E \), as Figure 5.3(b) illustrates. Once done, opening \( S_E \) and closing de-energizing switch \( S_D \), after a dead time where both are open, connects switching node \( v_{SW} \) to ground, which de-energizes L into \( C_{VAR} \). After \( C_{VAR} \) absorbs and exhausts L’s energy, capacitor voltage \( v_C \) reaches \( V_{BAT} \) and \( S_D \) disengages, which is when the precharge phase terminates. Both \( S_E \) and \( S_D \) remain off during any of the subsequent phases to avoid discharging \( C_{VAR} \). Note that fully draining L and allowing its current \( i_L \) to remain at zero for a finite fraction of the vibration period is analogous to operating L in discontinuous-conduction mode (DCM), when referring to switching converters.
While energizing, the precharge circuit in Figure 5.3 can be described as a differential equation, which, assuming both \( v_C \) and \( i_L \) (which equals \( i_C \) and \( i_{BAT} \)) are initially zero, results in the following (as derived in Appendix A):

\[
v_C(t) = V_{BAT}[1 - \cos(\omega_{LC}t)]
\]

and

\[
i_L(t) = i_C(t) = i_{BAT}(t) = V_{BAT} \sqrt{\frac{C_{MAX}}{L}} \sin(\omega_{LC}t),
\]

where \( \omega_{LC} \) is the LC’s natural resonant frequency. The battery energizes \( L \) and \( C_{MAX} \), where the stored energy in both components equals

\[
E_{LC}(t) = \frac{1}{2} L i_L^2(t) + \frac{1}{2} C_{MAX} v_C^2(t),
\]

which simplifies to

\[
E_{LC}(t) = C_{MAX} V_{BAT}^2[1 - \cos(\omega_{LC}t)].
\]

As a result, \( E_{LC} \) reaches \( E_{INV} \) in

\[
t_e = \frac{\pi}{3} \sqrt{\frac{LC_{MAX}}{L}},
\]

which corresponds to a \( C_{VAR} \) target voltage of 0.5\( V_{BAT} \). In other words, \( S_E \) should engage and allow the battery to energize \( L \) and \( C_{VAR} \) until \( C_{VAR} \) charges to 0.5\( V_{BAT} \), after which point \( S_D \) should allow \( L \) to finish charging \( C_{VAR} \) to its target of \( V_{BAT} \). After energizing \( L \), its peak current is

\[
i_{L,PK} = \frac{V_{BAT}}{2} \sqrt{\frac{3C_{MAX}}{L}}.
\]

Note that although the energy invested by the battery (\( E_{INV} \)) equals the energy received in \( C_{VAR} \) (ideally), the total charge lost by the battery does not equal the charge gained in \( C_{VAR} \). This difference arises because of the voltage inequality between the two, as
drawing power from a larger voltage (e.g., $V_{\text{BAT}}$ is greater than the initial voltage of $v_C$) requires less current. The total charge collected in $C_{\text{VAR}}$ ($\Delta q_C$) as it charges from zero, for instance, is

$$\Delta q_C = C_{\text{MAX}} V_{\text{BAT}},$$

whereas the charge lost by the battery ($\Delta q_{\text{BAT}}$) is

$$\Delta q_{\text{BAT}} = \int i_{\text{BAT}}(t) dt = \frac{1}{2} C_{\text{MAX}} V_{\text{BAT}},$$

which is half the final charge in $C_{\text{VAR}}$. In practice, however, power losses across the system, delays, and other non-idealities when energizing and de-energizing $L$ dissipate a portion of $E_{\text{INV}}$, which means $C_{\text{VAR}}$’s energizing target voltage must exceed $0.5V_{\text{BAT}}$ to compensate for these losses. In other words, $V_{\text{BAT}}$ must overinvest energy to overcome losses in the circuit.

Once $C_{\text{VAR}}$ and $L$ store $E_{\text{INV}}$, they are disconnected from $V_{\text{BAT}}$, and $L$ de-energizes to $C_{\text{VAR}}$. Capacitor voltage $v_C$ and inductor current $i_L$ (which is the same as $i_C$), as derived in Appendix A, are

$$v_C(t) = \frac{1}{2} V_{\text{BAT}} \cos(\omega_{Lc} t) + I_{L, pk} \frac{L}{C_{\text{MAX}}} \sin(\omega_{Lc} t)$$

and

$$i_L(t) = i_C(t) = I_{L, pk} \cos(\omega_{Lc} t) - \frac{1}{2} V_{\text{BAT}} \frac{C_{\text{MAX}}}{L} \sin(\omega_{Lc} t),$$

where now the initial capacitor voltage is $0.5V_{\text{BAT}}$. $E_{\text{INV}}$ transfers completely to $C_{\text{VAR}}$ when $v_C$ reaches $V_{\text{BAT}}$, and therefore, $i_L$ drops to zero. Note that an asynchronous diode could replace $S_D$ to simplify control since the diode automatically turns on when current
flows through it, but at the expense of higher conduction losses due to its forward bias voltage. As a result, de-energizing L lasts

\[ t_D = \frac{\pi}{3} \sqrt{LC_{\text{MAX}}}, \]  

(5.22)

and the total precharge time, ideally, is

\[ t_{\text{PCH}} = t_E + t_D = \frac{2\pi}{3} \sqrt{LC_{\text{MAX}}}. \]  

(5.23)

5.1.3 Comparison

Both inductor-based precharge circuits presented in Figures 5.2 and 5.3 transfer \( E_{\text{INV}} \) from the battery to charge \( C_{\text{VAR}} \) to \( V_{\text{BAT}} \) efficiently. Although in theory, the circuits are lossless, in a practical implementation they do dissipate energy, although significantly less than the switch-based precharger in Figure 5.1 because the inductor allows the voltages across the switches to remain low (in the mV range) while they conduct \( i_L \). For instance, the voltages across the switches are small but not zero, inducing finite conduction losses, and parasitic capacitors present in the circuit (such as the gate capacitance of each transistor switch) require energy to charge and discharge. This is the reason why the two-switch precharger from Figure 5.3 is preferred, since less switches lead to fewer losses, simpler control, and easier integration into smaller silicon area. Yet, it is important to note that the four-switch solution permits recovery of the remnant energy (at \( C_{\text{MIN}} \)), if viable and desired, through the same inductor.

The duration of each energy transfer lasts approximately 100-300 ns, significantly shorter than typical vibration periods, which are in milliseconds, as for example, 33.3 ms for 30 Hz vibrations. As a result, \( C_{\text{VAR}} \)’s vibration-induced variation, as perceived by the precharger, is slow enough to seem constant near \( C_{\text{MAX}} \). Note that each precharger
energizes and de-energizes L once every vibration cycle, allowing its current $i_L$ to remain at zero for the rest of the vibration period, analogous to a non-inverting buck-boost (four-switch version) or buck (two-switch version) converter in deep discontinuous-conduction mode.

### 5.2 Harvesting Switch

#### 5.2.1 Asynchronous Diode Switch

When harvesting, vibrations separate $C_{VAR}$’s parallel plates (i.e., $C_{VAR}$ decreases) and, since $V_{BAT}$ clamps its voltage though ideal switch $S_H$, seen in Figure 5.4(a), drive charge $q_{HARV}$ (in the form of $i_{HARV}$) and energy $E_{HARV}$ into the battery. The simplest embodiment of the switch is an asynchronous diode because no additional circuit is required to control it, as it automatically conducts the harvesting current to the battery when available and blocks reverse current when $C_{VAR}$ reaches $C_{MIN}$ (since a diode is unidirectional). The drawback is its forward voltage drop $v_D$, which implies no current flows until $v_C$ rises from precharged voltage to $V_{BAT} + v_D$. This corresponds to $C_{VAR}$ decreasing from $C_{MAX}$ to $C_{MAX,D}$ under charge-constrained conditions. In other words, the total capacitance variation reduces from $\Delta C_{VAR}$ to $\Delta C_{VAR,D}$ or
\[ \Delta C_{\text{VAR},D} = C_{\text{MAX,D}} - C_{\text{MIN}} = \left( \frac{V_{C,\text{FINAL}}}{V_{\text{BAT}} + v_D} \right) C_{\text{MAX}} - C_{\text{MIN}}, \]  

(5.24)

where \( V_{C,\text{FINAL}} \) is the final precharge voltage of \( C_{\text{VAR}} \). Some of the converted energy then is diverted from the battery to overcome \( v_D \). This remains true even when considering a higher \( v_C \) induces a higher harvesting current \( (i_{\text{HARV,D}}) \) because harvesting time \( t_{\text{HARV,D}} \) is now shorter by the length of time \( C_{\text{VAR}} \) takes to reach \( C_{\text{MAX,D}} \):

\[ i_{\text{HARV,D}} = (V_{\text{BAT}} + v_D) \frac{\Delta C_{\text{VAR},D}}{t_{\text{HARV,D}}}. \]  

(5.25)

As a result, \( v_D \) reduces harvested energy gain \( (E_{\text{HARV,D}}) \) to

\[ E_{\text{HARV,D}} = V_{\text{BAT}} i_{\text{HARV,D}} t_{\text{HARV,D}} = V_{\text{BAT}} (V_{\text{BAT}} + v_D) \Delta C_{\text{VAR},D}. \]  

(5.26)

Precharging \( C_{\text{VAR}} \) to \( V_{\text{BAT}} + v_D \) circumvents the brief charge-constrained event mentioned and recovers the full \( C_{\text{VAR}} \) variation, but requires a higher energy investment \( E_{\text{INV,D}} \) from the battery. The problem is the optimum precharge voltage of the system is \( V_{\text{BAT}} \) so \( V_{\text{BAT}} + v_D \) produces a less-than-optimal net energy gain per cycle \( E_{\text{NET,D}} \). For proof, consider that differentiating \( E_{\text{NET,D}} \) or

\[ E_{\text{NET,D}} = E_{\text{HARV,D}} - E_{\text{INV}} = E_{\text{HARV,D}} - \frac{1}{2} C_{\text{MAX}} V_{C,\text{FINAL}}^2 \]  

(5.27)

with respect to final precharge voltage \( V_{C,\text{FINAL}} \), equating to zero, and solving for \( V_{C,\text{FINAL}} \) yields \( V_{\text{BAT}} \) as optimum

\[ \frac{\partial E_{\text{NET,D}}}{\partial V_{C,\text{FINAL}}} = 0 \rightarrow V_{C,\text{FINAL,OPT}} = V_{\text{BAT}}. \]  

(5.28)

A diode therefore produces a non-optimal net energy gain of

\[ E_{\text{NET,D}} = E_{\text{NET}} - V_{\text{BAT}} v_D C_{\text{MIN}}, \]  

(5.29)
which compared to an ideal switch case \( E_{\text{INV}} \) in Equation 4.16, yields lower energy. As an example, a 400-100 pF \( C_{\text{VAR}} \) variation harvests in simulations 3.45 nJ and 3.69 nJ with and without the diode, respectively, where the 0.7 V drop decreased \( C_{\text{MAX}} \) from 400 pF to about 330 pF, reducing energy by 240 pJ/cycle (by 6.5%), which is significant considering how difficult decreasing losses at these low power levels is.

### 5.2.2 Synchronous MOS Switch

A synchronous transistor switch, with respect to its voltage drop, more closely resembles an ideal switch. A transistor, however, is bidirectional, and requires additional circuitry to synchronize with the vibrations that cause capacitance change. For instance, a control block must detect when to close the transistor at the start of harvesting and when to open it at the end of the phase. As a result, it requires additional circuits, and therefore power, to control it and drive the parasitic capacitors it presents. If a single PMOS transistor implements \( S_{\text{H}} \), its n-well backgate (shorted to the transistor source) must be connected to \( V_{\text{BAT}} \) to ensure that the parasitic pn-junction diode (between p-type drain
implant and n-well) is reversed biased when $C_{VAR}$ is discharged ($v_C < V_{BAT}$). Otherwise, the lossy body diode would forward bias and bypass the inductor-based pre-charger, directly charging $C_{VAR}$, at the expense of large energy losses. Nevertheless, $C_{VAR}$ could be overcharged to the point where $v_C$ exceeds $V_{BAT}$ enough to turn on the parasitic p-n junction, and return the excess energy to the battery. The problem is that the current conduction could forward bias the parasitic pnp bipolar-junction transistor (BJT) present in PMOS devices in standard, n-well, p-substrate technologies (between the drain p-implant, n-well, and p-substrate) and, as a result, current diverts to ground, rather than returning to the battery. An n-buried layer (NBL) below the n-well, as well as careful layout, mitigates the effects of the parasitic BJT and discourages substrate currents.

A safer solution is to place two PMOS transistors in series, as $MP_{HA}$ and $MP_{HB}$ shown in Figure 5.6. Both transistors effectively constitute single switch $MP_H$ but are separate to ensure that their parasitic body diodes and BJTs, connected back-to-back, do not forward bias or turn on during the precharge phase (even if $C_{VAR}$ is overcharged), where each device blocks the other’s body diode from conducting when disengaged. $C_{VAR}$ must be precharged as close to $V_{BAT}$ as possible to minimize losses across the switch, which are proportional to the voltage difference between $V_{BAT}$ and $V_{C,FINAL}$ as Equation 5.3 shows. In other words, undercharging $C_{VAR}$ results in $V_{BAT}$ finishing
precharge through the lossy switch, while overcharging $C_{VAR}$ returns excess energy to $V_{BAT}$, but part of it is lost in the switch and precharger circuit. Using the synchronous switch merits adoption, but only if the control circuit dissipates sufficiently less energy than what the diode effectively loses, which is why biasing control circuits in subthreshold is important.

### 5.3 Power Losses of Proposed Energy Harvester

The energy harvester to be presented in subsequent chapters minimizes losses by precharging $C_{VAR}$ with a two-switch inductor-based circuit and connecting it to $V_{BAT}$ during harvesting through back-to-back PMOS transistors, as shown in Figure 5.7. Resistances and body diodes determine the conduction power losses of the precharger circuit. Inductor current $i_L$ flows through the cumulative equivalent series resistance (ESR) of the capacitor ($R_{ESR,C}$), inductor ($R_{ESR,L}$), and battery ($R_{ESR,B}$), and the channel...
impedances of both switches, $MPE$ and $MND$, which are a function of their width-to-length ratio and gate voltages. The resulting resistive power losses are

$$P_{\text{COND,PCH}} = I_{L,RMS}^2 R_{\text{EQ}} t_{\text{PCH}} f_{\text{VIB}},$$

where $P_{\text{COND,PCH}}$ is averaged over time, $I_{L,RMS}$ is the root-mean square (RMS) value of the approximately triangular inductor current, $R_{\text{EQ}}$ the equivalent resistance of the circuit path in question, $t_{\text{PCH}}$ the total conduction time, and $f_{\text{VIB}}$ the vibration frequency. $I_{L,RMS}$ always flows through $R_{\text{ESR,L}}$, $R_{\text{ESR,C}}$, and: (1) when energizing $L$, $R_{\text{ESR,B}}$ and $r_{DS,MPE}$; and (2) when de-energizing to $C_{\text{VAR}}$, $r_{DS,MND}$. By representing $i_L$ as a triangular function with peak $I_{L,PK}$ that lasts $t_{\text{PCH}}$, the conduction losses are approximately

$$P_{\text{COND,PCH}} = N \frac{1}{3} I_{L,PK}^2 (r_{\text{DS}} + R_{\text{ESR,L}} + R_{\text{ESR,BC}}) t_{\text{PCH}} f_{\text{VIB}},$$

where $N$ is the number of inductor storage-delivery cycles within the precharge phase, $I_{L,PK}$ the peak inductor current, and $R_{\text{ESR,BC}}$ represents the ESR of the battery and $C_{\text{VAR}}$. During the dead time inserted between energize and de-energize steps, current flows through the body diode of $MND$, $R_{\text{ESR,L}}$, and $R_{\text{ESR,C}}$, incurring additional conduction power losses

$$P_{\text{COND,DEAD}} = N \left[ I_{L,PK}^2 (R_{\text{ESR,L}} + R_{\text{ESR,C}}) + I_{L,PK} V_D \right] t_{\text{DEAD}} f_{\text{VIB}},$$

where the dead time current is approximately $I_{L,PK}$, $V_D$ is the voltage drop across the body diode, and $t_{\text{DEAD}}$ is the dead time.

At each switching event, each transistor conducts current before fully turning on or off, resulting in significant drain-source voltage to overlap with channel current and incurring power losses. For instance, when initially engaged, the drain-source voltage of $MPE$ is $V_{\text{BAT}}$, but its current is low since $L$ has not energized yet, and overlap losses are
not significant. When MP_E shuts off, however, the peak inductor current flows through the transistor until its drain-source voltage increases to $V_{BAT}$, at which point the body diode of MN_D conducts current, potentially causing during this brief time considerable overlap losses. By assuming that the time that current and voltage overlap $t_{OV}$ is dominated by the charge time of MP_E’s gate-drain capacitance and that the average drain-source voltage during this time is half of its peak value ($0.5V_{BAT}$), the average overlap power lost is approximately

$$P_{OV} = N \frac{1}{2} V_{BAT} I_{L,PK} t_{OV} f_{VIB}.$$  \hfill (5.33)

On the other hand, when switch MN_D engages, its drain-source voltage is always relatively low since there is only its body diode drop across its channel, and, when it shuts off, $i_L$ is near zero. Therefore, overlap losses of MN_D are trivial.

Each transistor features parasitic gate capacitances that require energy to charge and discharge during each switching event. These losses are typically referred to as gate-drive power losses and the average power lost per precharge event per switch is

$$P_{GD} = NC_G V_{DRV}^2 f_{VIB},$$  \hfill (5.34)

where $C_G$ is the equivalent parasitic capacitance at the device gate and $V_{DRV}$ is the gate drive voltage, which is in this case equal to $V_{BAT}$. The driving stage of each transistor incurs these losses. Substituting MN_D with a diode eliminates the gate-drive requirements of the transistor, but at the expense of significantly higher conduction losses because of the diode forward voltage drop. Also, note that even though energizing L in several steps reduces the peak inductor current and therefore conduction losses, it results in greater gate-drive losses since more switching events occur. Adopting smaller gate area transistors reduces the parasitic capacitance effects of switching losses such as gate-drive
and overlap losses, and, accordingly, \( \text{MP}_{E} \) and \( \text{MN}_{D} \) feature minimum gate lengths. Since gate widths determine the effective channel resistances of the switches, however, their values represent a compromise between conduction and switching losses. Finally, note that \( V_{\text{BAT}} \) also charges other parasitic capacitances present in the system, such as \( C_{\text{PAR}} \) and \( C_{\text{SW}} \) in Figure 5.7, resulting in higher energy investment.

While harvesting, only transistors \( \text{MP}_{HA} \) and \( \text{MP}_{HB} \) interact directly with the harvesting current. The transistors feature minimum size gate length and width so that their parasitic capacitances do not create charge leakage problems and considerable gate-driving losses. However, their cumulative channel resistance \( (r_{\text{DS,H}}) \) is relatively larger and the resulting conductive power loss is

\[
P_{\text{HARV, PCH}} = I_{\text{HARV, AVG}}^2 (r_{\text{DS,H}} + R_{\text{ESR,BC}}) t_{\text{HARV}} f_{\text{VIB}},
\]

where \( I_{\text{HARV, AVG}} \) is the average harvesting current and \( t_{\text{HARV}} \) represents the time it takes the capacitor to reach its minimum capacitance point. The actual constrained voltage is the sum of \( V_{\text{BAT}} \) and the switch drain-source voltage (i.e., \( r_{\text{DS,H}} i_{\text{HARV}} \)), and hence, the electrostatic force against which the mechanical accelerations work must therefore increase to raise \( v_{C} \) slightly above the battery by the Ohmic drop of the switch resistance. As a result, the power dissipated by the switch is provided by extra work performed by the mechanical device and not from the battery or the electrical energy generated. However, this assumes that the design of the device took into account the extra force on the capacitor by reducing its stiffness, or otherwise, the total capacitance variation would decrease.

At the end of precharge, \( \text{MP}_{H} \) connects \( C_{\text{VAR}} \) to \( V_{\text{BAT}} \), and any difference between final precharged voltage \( V_{C,\text{FINAL}} \) and the battery results in energy exchanged through the
lossy switch. As a result, because of the switch resistance, conduction losses result when either $V_{\text{BAT}}$ finishes precharging $C_{\text{VAR}}$ or $C_{\text{VAR}}$ returns overcharge to $V_{\text{BAT}}$. This power loss is approximately

$$P_{\text{SW}} = E_{\text{SW}} f_{\text{VIB}} = \frac{1}{2} C_{\text{MAX}} (V_{\text{BAT}} - V_{C,\text{FINAL}})^2 f_{\text{VIB}},$$

(5.36)

which is the energy lost per cycle from Equation 5.3 times the vibration frequency. To minimize conduction losses across the harvesting switch, it is important to ensure that the precharge control circuits charge $C_{\text{VAR}}$ as close to $V_{\text{BAT}}$ as possible.

### 5.4 Simulation Results

Increasing the energy gain of an electrostatic voltage-constrained energy harvester amounts to understanding its energy budget and reducing the losses associated with each operational phase. Using an inductor-based precharger, for example, increases the precharger efficiency from 50\% to 79.3\%, as the simulation results shown in Figure 5.8 demonstrate. Simulations show how a 3.5-V battery invests 2.75 nJ with a peak current of
14.6 mA through 10 µH to precharge 391.4 pF from 1.05 V (which had 0.22 nJ stored from the previous reset phase), even in the presence of 10-pF and 4-pF parasitic capacitances at $v_C$ and $v_{SW}$ nodes, respectively.

When connected to $V_{BAT}$ for harvesting, Figure 5.9 shows that the simulated 1.3-kHz vibrations generate up to 4.48 µA that harvest 3.65 nJ as $C_{VAR}$ decreases to 100 pF, saving up to 200 pJ/cycle by otherwise using a diode. During reset in Figure 5.10, as

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**Figure 5.9.** Simulations showing energy harvesting with a PMOS-based switch.

**Figure 5.10.** Simulations showing capacitor voltage $v_C$ as $C_{VAR}$ increases during the reset phase.
capacitance increases again from $C_{\text{MIN}}$, $v_C$ slowly drops to 1.05 V, at which point $C_{\text{MAX}}$ is detected and the subsequent precharge phase begins. The remnant energy left in $C_{\text{VAR}}$ after the harvesting phase is used to sense its capacitance and, therefore, reduce circuit complexity and control power. Each of these strategies make it possible to maintain a net energy gain in the midst of all the losses that exists in any practical circuit, where harvested energy is greater than what is invested and lost, resulting in a net energy gain of 900 pJ/cycle for 400 to 100 pF variations.

5.5 Summary

As part of the battery-constrained energy-harvesting scheme presented in Chapter 4, $C_{\text{VAR}}$ must be precharged to $V_{\text{BAT}}$ when capacitance is at $C_{\text{MAX}}$. To do this efficiently and to avoid losses from exceeding the harvested energy, a two-switch inductor-based precharger transfers the required investment from the battery to $C_{\text{VAR}}$. In addition, during the harvesting phase, current $i_{\text{HARV}}$ channels directly to $V_{\text{BAT}}$ through a transistor-based switch that prevents the losses a diode might otherwise incur. In these ways, by precharging $C_{\text{VAR}}$ with an inductor-based circuit and connecting it to $V_{\text{BAT}}$ during harvesting through a transistor switch, the energy harvester limits extraneous losses and helps preserve a net energy gain, which is the fundamental challenge of any energy-harvesting system.

Charging the battery with ambient energy still requires monitoring, control, and precharge circuits that demand power to function. In practice, $E_{\text{HARV}}$ must overcome not only precharge investment $E_{\text{INV}}$ but also the losses associated with each phase in the cycle. This is a considerable challenge because $E_{\text{HARV}}$ is low to begin with and each phase requires monitoring and control circuitry to detect $C_{\text{MAX}}$ during reset to initiate the
precharge phase and $C_{\text{MIN}}$ to disconnect $C_{\text{VAR}}$ from the battery at the end of the harvesting phase. To mitigate these quiescent losses incurred by the additional control circuitry, the presented system employs further low-energy strategies such as operating with subthreshold currents and shutting off unused components.
The basic principle of the battery-constrained harvester integrated circuit (IC), which the next two chapters will discuss, relies on clamping $C_{\text{VAR}}$’s voltage $v_C$ with the microsystem’s battery. Vibrations then work against the variable capacitor’s electrostatic force to separate its plates, decreasing $C_{\text{VAR}}$ from $C_{\text{MAX}}$ to $C_{\text{MIN}}$, and converting ambient kinetic energy to electrical in the form of harvesting current $i_{\text{HARV}}$. Before connecting $C_{\text{VAR}}$ to the battery, however, the system precharges $C_{\text{VAR}}$ (when at $C_{\text{MAX}}$) with a quasi-lossless inductor-based circuit to battery voltage $V_{\text{BAT}}$ by investing the necessary energy to establish the required electrostatic force for vibrations to work against. The inductor-based precharger avoids incurring considerable conduction losses in the connecting switch, which otherwise would dissipate much of the energy harvested. When $C_{\text{VAR}}$ connects to the battery, vibrations generate $i_{\text{HARV}}$, which directly charges the battery and augments its energy. Subsequently, after reaching $C_{\text{MIN}}$, $C_{\text{VAR}}$ disconnects to avoid the reverse process from discharging the battery, leaving $C_{\text{VAR}}$ open-circuited instead so that $v_C$ decreases and resets to a lower value as $C_{\text{VAR}}$ returns to $C_{\text{MAX}}$ again and the next cycle begins.

For the energy in the battery to increase with each vibration cycle, harvested energy must not only surpass what the battery invests, but also the losses associated with the monitoring and control circuits associated with each phase in the energy-harvesting cycle. Therefore, the IC to be presented in this chapter shuts off components when unused and operates with subthreshold currents when necessary. Consequently, the
precharger, harvester, monitoring, and control microelectronics draw enough power to operate, yet allow the system to yield a net energy gain.

### 6.1 System

The proposed IC coordinates each phase in the energy-harvesting cycle by (i) detecting $C_{\text{MAX}}$ during reset, (ii) precharging $C_{\text{VAR}}$, (iii) connecting $C_{\text{VAR}}$ to the battery to allow vibrations to drive $i_{\text{HARV}}$ into the battery, (iv) detecting $C_{\text{MIN}}$ during the harvesting phase, and (v) disconnecting $C_{\text{VAR}}$ from everything during reset. Each functional block in the system, shown in Figure 6.1, corresponds to a phase in the process that the digital controller enables and powers in sequence, one at a time as each phase occurs. For instance, the precharge detection block monitors when $C_{\text{VAR}}$ reaches $C_{\text{MAX}}$ during reset to trigger the next phase in the cycle, which is precharge. During precharge, the detection circuits shut off and the IC charges $C_{\text{VAR}}$ to $V_{\text{BAT}}$ through inductor $L$. Afterwards, the
precharger circuits shut off, prompting harvesting switch $S_H$ to connect $C_{VAR}$ to $V_{BAT}$, allowing $i_{HARV}$ to charge the battery during the ensuing harvesting phase. During this phase, the controller powers the harvest detection circuits so they monitor when $C_{VAR}$ reaches $C_{MIN}$ to break the $C_{VAR}$-$V_{BAT}$ connection, after which $C_{VAR}$ resets and the controller again enables the precharger, allowing the cycle to repeat as vibrations swing $C_{VAR}$ between $C_{MAX}$ and $C_{MIN}$ every 33.3 ms (at the target vibration frequency of 30 Hz).

To start, the quasi-lossless inductor-based switching circuit in Figure 6.1, which Chapter 5 discusses in detail, transfers $E_{INV}$ from the battery to charge $C_{VAR}$ to $V_{BAT}$. To this end, the battery energizes both $L$ and $C_{VAR}$ with $E_{INV}$, when at $C_{MAX}$, by closing energizing switch $S_E$. Once done, opening $S_E$ and closing switch $S_D$ de-energizes $L$ into $C_{VAR}$. After $C_{VAR}$ absorbs and exhausts $L$’s energy, capacitor voltage $v_C$ reaches $V_{BAT}$ and $S_D$ disengages, which is when the precharge phase terminates. Notice the reason the circuit is significantly more efficient than simply charging $C_{VAR}$ through $S_H$ is because $L$ allows the voltages across the switches to remain low (in the mV range) while they conduct $i_L$. Otherwise, considerable voltage and current would coexist across $S_H$, resulting in losses that would dissipate any gains from the harvested energy from the ambient vibrations.

Ideally, to charge $C_{VAR}$ to $V_{BAT}$, the battery should energize $L$ and $C_{VAR}$ for time $t_E$ (Equation 5.16), which is one-sixth of their natural frequency $\omega_{LC}$, and corresponds to charging $C_{VAR}$ during the energizing phase to half of the battery voltage ($0.5V_{BAT}$). In practice, however, power losses across the system, delays, and other non-idealities dissipate a portion of $E_{INV}$, which means $C_{VAR}$’s energizing target voltage, which is set by reference voltage $V_{REF}$ in Figure 6.1, must exceed $0.5V_{BAT}$ to compensate for these
losses. In other words, actual $E_{\text{INV}}$ must exceed its theoretical lower bound. Then, after fully de-energizing $L$ into $C_{\text{VAR}}$ when $i_L$ approaches zero, $S_D$ shuts off to avoid discharging $C_{\text{VAR}}$. Note the precharge phase only lasts a small fraction of the vibration period (about 100–300 ns of 0.01–1 s). In other words, precharge is drastically faster than vibrations, which means that to the precharger, $C_{\text{VAR}}$ seems constant near $C_{\text{MAX}}$.

After precharge, the IC connects $C_{\text{VAR}}$ to $V_{\text{BAT}}$ and vibrations decrease $C_{\text{VAR}}$ and generate $i_{\text{HARV}}$, which charges the battery. The controller disconnects $C_{\text{VAR}}$ once it reaches $C_{\text{MIN}}$ because $i_{\text{HARV}}$ in reverse would otherwise discharge the battery. Due to the intrinsic resistance of connecting switch $S_H$, $i_{\text{HARV}}$ induces a voltage drop that forces $C_{\text{VAR}}$ to raise its voltage $v_C$ slightly above $V_{\text{BAT}}$ during the harvesting phase. The harvest-detection block then monitors $v_C$ and prompts $S_H$ to disengage when $v_C$ drops to $V_{\text{BAT}}$, which happens when $i_{\text{HARV}}$ decreases to zero, that is, when $C_{\text{VAR}}$ reaches $C_{\text{MIN}}$. Although a diode would engage and disengage automatically with $i_{\text{HARV}}$ asynchronously, its forward voltage drop requires a brief charge-constrained phase as $v_C$ increases a diode voltage above $V_{\text{BAT}}$, resulting in energy losses, as previously discussed in Chapter 5. Using synchronous switch $S_H$ is more efficient than a diode as long as its control circuitry consumes less power than the diode. For optimal results, therefore, currents in the nanoampere range bias the harvesting detection block in subthreshold and the system’s control logic enables it only during the harvesting phase, which is roughly half of each vibration cycle.

After harvesting, the IC disconnects $C_{\text{VAR}}$ and leaves it open-circuited during the reset phase so $v_C$ can decrease automatically when $C_{\text{VAR}}$ increases from $C_{\text{MIN}}$ to $C_{\text{MAX}}$. During this phase, the precharge detector indirectly senses $C_{\text{VAR}}$ and detects when it
reaches $C_{\text{MAX}}$. Because $v_C$ decreases with increasing $C_{\text{VAR}}$, $v_C$ reaches its minimum value $V_{C_{\text{MIN}}}$ and begins to increase after $C_{\text{VAR}}$ reaches $C_{\text{MAX}}$ and starts to decrease. This means $V_{C_{\text{MIN}}}$ corresponds to when $C_{\text{VAR}}$ is at $C_{\text{MAX}}$. In this way, the IC uses the remnant energy left in $C_{\text{VAR}}$ after the harvesting phase to detect $C_{\text{MAX}}$, which the system would otherwise lose. Like the harvesting counterpart, subthreshold currents bias the precharge detector and it only engages during reset, which is half the vibration cycle.

Figure 6.2. Prototype energy harvester IC (transistor dimensions are in $\mu$m).
6.2 Integrated Circuit

The complete harvesting system, which Figure 6.2 shows, integrates all blocks into a single silicon IC, with the exception of $L$, $C_{VAR}$, and the current-setting resistors of the nanoampere bias-current generator. Resistors $R_{DLYA}$ and $R_{HARV}$ are off chip for testability purposes only, to freely adjust their values when experimenting with the IC. Similarly, the IC relied on off-chip reference voltage $V_{REF}$ to modify the precharging target voltage easily during experiments, optimum values of which depend on the losses across the system.

6.2.1 Precharger

Transistors $M_{PE}$ and $M_{ND}$ in Figure 6.3 and their inverting drivers energize and de-energize $L$ to charge $C_{VAR}$ to $V_{BAT}$. As $L$ and $C_{VAR}$ energize, comparator $C_{PC}$ senses $v_C$ until it reaches reference voltage $V_{REF}$, at which point $C_{PC}$ opens $M_{PE}$ and, after a dead period during which time both switches remain off, $M_{ND}$ closes to de-energize $L$.

![Figure 6.3. Precharger circuit (transistor dimensions are in μm).](image)
Essentially, $V_{\text{REF}}$ sets energizing time $t_E$ and, therefore, the invested energy and the final precharge voltage. The inductor’s de-energizing current $i_L$ then flows to $v_C$ through $M_{ND}$, causing $v_{SW}$ to fall slightly below 0 V since $i_L$ flows from ground. As $i_L$ decreases, $v_{SW}$ increases from below ground towards 0 V until $i_L$ falls to 0 A. Comparator $C_{P_{SW}}$ senses when $i_L$ reaches 0 A indirectly by monitoring when $v_{SW}$ rises to 0 V, which indicates the end of the de-energizing step and the precharge phase.

Nevertheless, comparator $C_{P_{SW}}$’s delay would keep $M_{ND}$ engaged long enough to discharge $C_{VAR}$, had not a built-in offset voltage $V_{OS}$ been included. The offset shifts the trip point so that $C_{P_{SW}}$ starts tripping just before $v_{SW}$ reaches 0 V, relying on its delay for $M_{ND}$’s gate voltage to transition and shut off when $v_{SW}$ actually nears 0 V. Before either $C_{P_{VC}}$ or $C_{P_{SW}}$ becomes functional, however, the first step in the precharge process is to power their local bias-current generator. Once the bias is ready, $M_{PE}$ closes and the generator powers both comparators, but only enables $C_{P_{VC}}$, forcing $C_{P_{SW}}$ to remain high until the start of the de-energizing step. Note precharge only occurs during a diminutive fraction of the entire cycle (e.g., about 200 ns of 33 ms – that is less than 0.001%) so all precharge circuits must be sufficiently fast, which means transistors in $C_{P_{VC}}$ and $C_{P_{SW}}$ must operate in strong inversion to feature sufficiently short delays in the nanosecond range. Even so, because their currents only flow during precharge, they do not represent a significant energy loss.

Each power transistor $M_{PE}$ and $M_{ND}$ features minimum gate lengths to mitigate switching losses due to their gate capacitance. Their gate widths, however, represent a compromise between conduction and switching losses, where smaller widths lead smaller gate capacitance at the expense of greater drain-source channel resistance and, therefore,
greater conduction losses. Inverting gate-drivers \( GD_P \) and \( GD_N \) drive \( MP_E \)’s and \( MN_D \)’s gates, respectively, where, to save energy, each driving circuit only features three series inverting stages that progressively increase in size, as seen in Figure 6.4. In this manner, digital signals \( v_{DRVP} \) and \( v_{DRVN} \) from control logic are augmented to drive the relatively large capacitive load the gate of each power switch presents.

As previously discussed, \( CP_{VC} \) signals the end of the energizing step when \( v_C \) reaches \( V_{REF} \). Comparator \( CP_{VC} \) in Figure 6.5 uses an n-type input pair with a p-type load mirror to feed a common-source transistor and subsequently drive a digital inverter. High-impedance cascode current sources bias each gain stage and all the NMOS and PMOS bulks connect to ground and \( V_{BAT} \), respectively, unless otherwise specified. Common-source amplifier \( MP_2 \) further amplifies the signal from the first stage to decrease shoot-through (short-circuit) current in the ensuing inverters because a steeper
transition decreases the time pull-up and -down transistors conduct simultaneously. Switch MNH sinks additional current from the current mirror when VO2 is high to establish hysteresis by unbalancing the currents flowing through current-mirror load MP11-MP12. The differential pair features minimum channel lengths to keep delays short, even at the expense of accuracy, given adjustments in VREF can compensate for offsets.

The comparator generates fast and slow outputs vE-END(Fast) and vE-END(Slow) to create dead time between the energizing and de-energizing steps that prevent shoot-through (short-circuit) currents in the precharger power stage. When CPVC trips high, for example, vE-END(Fast) first opens PE to end the energizing step and, after the short delay RSL and CSL produce, vE-END(Slow) closes MN0 to start the de-energizing counterpart and enables CPSW. Note only one decision in CPVC, as in the other comparators in the system, matters: when its output transitions high. For this reason, class-A MP2’s sourcing current
sets this high-speed transition, while bias $I_{B2}$ slews the other slower transition, which
never occurs since the comparator disables after precharge.

Comparator $C_{P_{SW}}$’s input stage in Figure 6.6 monitors voltages near and below 0
V with a common-gate, gate-coupled NMOS differential pair. PMOS source followers
then level-shift the signal higher above ground to drive an NMOS differential pair, whose
output drives common-source PMOS amplifier $M_{P_4}$. Offset current $I_{OS}$ establishes a
systematic imbalance that produces input-referred offset $V_{OS}$ in $C_{P_{SW}}$. Similarly, switch
$M_{N_{H}}$ steers offset current $I_{H}$ only when output $v_{O4}$ is high to establish hysteresis so
$C_{P_{SW}}$ cannot trip again until $v_{SW}$ is well below 0 V. The bulbs of the input NMOS transistors
are connected to their respective sources to prevent $v_{SW}$, which swings below 0 V, from
inducing considerable substrate current through $M_{N_{12}}$’s body diode. Fortunately, the
systematic offset $I_{OS}$ creates is, by design, large enough to overwhelm the random offset
that results from placing $M_{N_{11}}$ and $M_{N_{12}}$ in separate p-type isolation tanks. Recall that at
this point in the cycle, after $i_L$ drops to 0 A, $C_{P_{SW}}$ trips low and signals the end of the de-
energizing step and precharge phase, allowing the harvesting phase to begin while
shutting all precharge circuits off and returning them to their previous states.
Figure 6.6. Precharge de-energize comparator CP_{SW} (transistor dimensions are in µm).
To generate the microampere-range bias currents \( CP_{VC} \) and \( CP_{SW} \) require, the precharge block features its own local bias-current generator. The circuit, shown in Figure 6.7, creates a first-order temperature-compensated current \( I_B \) by combining \( MP_3 \)’s proportional-to-absolute-temperature (PTAT) current with \( MP_{C3} \)’s base-emitter defined current, which is complementary-to-absolute-temperature (CTAT). The circuit impresses the base-emitter voltage difference between NPN pair \( Q_1 \)-\( Q_2 \) across \( R_{PTAT} \) to generate a PTAT current when \( R_{PTAT} \)’s temperature coefficient is low. Similarly, impressing \( Q_1 \)’s base-emitter voltage across \( R_{CTAT} \) induces a CTAT current. Combining both generates \( I_B \), from which all precharge circuits draw their biasing current.

The control logic only powers and enables the current generator during the precharge phase, which again, is a miniscule fraction of the vibration cycle. When power-
down signal $v_{PWRD}$ is high, the generator is off because MN$_{C1}$ pulls Q$_1$-Q$_2$’s base terminals to ground, keeping $R_{PTAT}$ and $R_{CTAT}$’s currents at 0 A. In addition, $v_{PWRD}$ engages MN$_{PW5}$ to charge and initialize $C_{PW}$ to near $V_{BAT}$ since MN$_{PW4}$ is off. The purpose of charging $C_{PW}$ at the gate of MN$_{PW2}$ is to prompt momentarily a startup response by pulling current $I_{PW}$ from mirror MP$_{P1}$-MP$_{P2}$. In other words, when $v_{PWRD}$ first transitions low to power the circuit, MN$_{PW3}$ engages and $C_{PW}$’s initial voltage induces MN$_{PW2}$ to sink considerable current ($I_{PW}$), but only until the circuit starts, after which point MN$_{PW4}$ discharges $C_{PW}$ and shuts MN$_{PW2}$ off [201]. Even though increasing $I_{PW}$ accelerates the startup time, quiescent losses in the circuit remain the same since no quiescent current flows through the startup circuit before or after it powers, unlike other conventional techniques. To allow the bias generator to settle to its steady state, the monitor circuit in Figure 6.8 only signals the system the current generator is ready when a delayed version of the current generated in Figure 6.7, produced by capacitors $C_1$ and $C_2$, is able to discharge $C_3$ sufficiently to trip current-limited inverter MN$_{2}$-MP$_{2}$.

![Figure 6.8. Current-bias ready-state recognition circuit (transistor dimensions are in μm).](image)
The digital control logic, shown in Figure 6.9, synchronizes the energizing and de-energizing steps in the precharge phase. More specifically, \( v_{PCH} \) from the precharge detector prompts the system to commence precharge by signaling the current generator to start, by triggering \( v_{PWRD} \) low with latch \( SR_{PWD} \). Once the generator is ready, \( v_{I-RDY} \) transitions high and forces PMOS-gate signal \( v_{GP} \) to close \( MP_E \). When the energizing step ends, \( CP_{VC} \)’s fast output \( v_{E-END(Fast)} \) shuts \( MP_E \) off and slower output \( v_{E-END(Slow)} \), after a short delay, engages \( MN_D \) and sets latch \( SR_{SW} \) with NMOS-gate signal \( v_{GN} \), where \( SR_{SW} \)’s output enables \( CP_{SW} \). Tripping \( CP_{SW} \)’s output \( v_{D-END} \) low forces \( v_{GN} \) low, which opens \( MN_D \). Signals \( v_{D-END} \) and \( v_{GN} \) reset power-down latch \( SR_{PWD} \) to trigger \( v_{PWRD} \) high and shut all precharge circuits off.

![Figure 6.9](image)

*Figure 6.9. Precharge logic operating during the (a) energize step, (b) de-energize step, and (c) through the end of the precharge phase.*
The rising edge detector in Figure 6.10(a) senses low-to-high transitions by comparing with an AND gate a digital input signal with its inverted and delayed counterpart. An initial low state therefore enables the AND gate to detect a high transition as the delayed input remains high temporarily. Detecting the opposite edge amounts to inverting all signals so the falling edge detector in Figure 6.10(b) simply inverts the input of another rising edge detector. With $SR_{PWD}$’s output down, since $v_{GN}$ and $v_{GP}$ are already low and high, respectively, precharge ends as $v_{PCH-END}$ transitions to a high state, which signals the harvesting phase to begin.

6.2.2 Harvest Detection

After precharge, back-to-back transistors $MP_{HA}$ and $MP_{HB}$ in Figure 6.2 connect $C_{VAR}$ to the battery. Each transistor blocks the other’s body diode to avoid unwanted currents from flowing into or out of the battery, and both are minimum size to reduce the parasitic capacitances they introduce. Their cumulative channel resistance creates a voltage drop that raises $v_c$ above $V_{BAT}$ when $i_{HARV}$ flows into the battery. Harvest-detect comparator $CP_{HARV-DET}$ monitors when this voltage drops to 0 V, or, in other words, when $v_c$ equals $V_{BAT}$, which indicates when harvesting ends since $i_{HARV}$ reduces to 0 A when $C_{VAR}$ reaches $C_{MIN}$. Since $i_{HARV}$ is in the nanoampere range, the voltage difference
between \( v_C \) and \( V_{BAT} \) is low, so inserting an additional 100 k\( \Omega \) (\( R_{HARV} \)) in the path between \( CP_{HARV-DET} \)’s inputs increases this voltage and alleviates \( CP_{HARV-DET} \)’s gain and resolution requirements.

When precharge ends, digital signal \( v_{PCH-END} \) sets the SR latch in Figure 6.11 to prompt \( MP_{HA} \) and \( MP_{HB} \), by forcing \( v_{HARV} \) low, to connect \( C_{VAR} \) to the battery and enable \( CP_{HARV-DET} \). When \( v_C \) drops to \( V_{BAT} \), \( CP_{HARV-DET} \) resets the latch to force \( v_{HARV} \) high and both break the \( C_{VAR}-V_{BAT} \) connection and disable \( CP_{HARV-DET} \), which subsequently returns to its previous high state. Signal \( v_{HARV} \) then prompts the precharge detection logic to power and enable the precharge detector during the ensuing reset phase.

Comparator \( CP_{HARV-DET} \) operates for roughly each half cycle, that is, 16.7 ms out of every 33.3 ms period for 30 Hz vibrations, which is considerably longer than the precharge circuits. For this reason, subthreshold currents bias the comparator to minimize its use of energy. \( CP_{HARV-DET} \)’s propagation delay is therefore in the microsecond range, which, although significantly slower than \( CP_{VC} \) and \( CP_{SW} \), is still considerably shorter than typical vibration periods. Because \( CP_{HARV-DET} \)’s input common-mode voltage slightly surpasses \( V_{BAT} \), NMOS source followers in Figure 6.12 level-shift the input down before feeding them into an NMOS differential pair, whose output feeds cascaded common-source amplifiers \( MP_3, MN_4, \) and \( MP_5 \). \( CP_{HARV-DET} \) includes several gain stages.
Figure 6.12. Harvest-detect comparator CP_HARV-DET (transistor dimensions are in µm).

because it must discern a small differential input voltage because of the low nanoampere
levels of $i_{HARV}$. Like before, MN_H sinks additional current from mirror MP_21-MP_22 when
$v_{O3}$ is high to establish hysteresis and only one transition in the class-A amplifiers
matters: when $v_{HV-END}$ falls.

Because $v_{O22}$ might discharge before $v_{O21}$ during power-up and cause CP_HARV-DET
to glitch and trip prematurely, the deglitch circuit in Figure 6.13 holds $v_{O22}$ to $v_{BAT}$ while
other nodes discharge and enough current flows through current-mirror MP_21-MP_22.
While disabled, deglitch circuit output $v_{EN-DLY}$ is at 0 V, keeping $v_{O22}$ and $v_{HV-END}$ at
$V_{BAT}$. When current source $I_B$ powers the circuit, $C_{G1}$ holds $v_{GL}$ down and MN_G3 off until
MP_G1’s current, which exceeds MN_G2’s, raises $v_{GL}$ above MN_G3’s threshold and $v_{EN-DLY}$
transitions to release its hold on $v_{O22}$ and $v_{HV-END}$. After CP_HARV-DET detects when $i_{HARV}$
reaches 0 A, the logic disables CP_HARV-DET, forcing its output high to prepare it for the
next harvesting phase.
6.2.3 Precharge Detection

During reset, \( v_C \) decreases as \( C_{VAR} \) increases in charge-constrained fashion, so sensing when \( v_C \) reaches \( V_{C\text{MIN}} \) equates to detecting \( C_{\text{MAX}} \). Precharge-detect comparator \( CP_{\text{PCH-DET}} \) compares \( v_C \) with a delayed version of itself, \( v_{CDLY} \), which the RC delay in Figure 6.14 generates. As a result, as \( v_C \) decreases, \( v_{CDLY} \) remains higher than \( v_C \) and \( CP_{\text{PCH-DET}} \)’s output \( v_{PCH} \) remains low. When \( v_C \) reaches \( V_{C\text{MIN}} \) and starts its ascent, \( v_C \) surpasses \( v_{CDLY} \) and causes \( CP_{\text{PCH-DET}} \) to trip and transition \( v_{PCH} \) to a high state, signaling \( C_{VAR} \) is ready for precharge.

The challenge is typical vibration frequencies are in the 1–100 Hz range, and generating a discernable voltage across \( CP_{\text{PCH-DET}} \) requires a substantial delay between \( v_C \) and \( v_{CDLY} \). Additionally, delay resistors dissipate energy proportional to the energy transferred to delay poly-poly capacitor \( C_{DLY} \), which represents an overinvestment. In
other words, smaller $C_{\text{DLY}}$ values require and, therefore, dissipate less energy. For this reason, $C_{\text{DLY}}$ is relatively low at 2 pF, and to generate the delay necessary to detect $V_{C,\text{MIN}}$ at 30 Hz, $R_{\text{DLY}}$ is 42 MΩ. Of the 42 MΩ, 12 MΩ are on chip as $R_{\text{DLYB}}$ in a very-high sheet-resistivity poly-silicon strip and, for testing flexibility, 30 MΩ are off chip as $R_{\text{DLYA}}$ in a thick-film resistor. Resistor $R_{\text{DLYB}}$ occupied 90 x 120 µm² of silicon area so integrating the remainder would have demanded 360 x 110 µm². To save area, substituting $R_{\text{DLY}}$ with a subthreshold-operated transconductor is possible, but at the expense of additional power losses and design complexity.

To conserve energy, the circuit only powers $CP_{\text{PCH-DET}}$ during reset. Digital signal $v_{\text{HARV}}$, which controls harvesting PMOS switch $MP_{H}$ (in Figure 6.2), transitions high to (i) disconnect $C_{\text{VAR}}$ from the battery at the end of the harvesting phase (when $C_{\text{VAR}}$ reaches $C_{\text{MIN}}$) and (ii) set the set-reset (SR) latch of the precharge detector on its rising edge. The inverted output of the latch then powers and enables $CP_{\text{PCH-DET}}$, whose output $v_{\text{PCH}}$ remains low (because disabling $CP_{\text{PCH-DET}}$ forces $v_{\text{PCH}}$ low) until $v_{C}$ begins to increase. Once $CP_{\text{PCH-DET}}$ trips, the rising edge of $v_{\text{PCH}}$ resets the latch, whose output subsequently disables $CP_{\text{PCH-DET}}$ and starts the precharge phase.
Figure 6.15. Precharge-detect comparator CP_{PCH-Det} (transistor dimensions are in µm).

Similar to the harvest-detect comparator, CP_{PCH-Det} conserves energy by powering for roughly each half cycle during reset and biasing with subthreshold currents. Its topology, shown in Figure 6.15, is similar to CP_{VC} in Figure 6.5, but without RC delay and a PMOS input pair to detect v_C as it drops during reset. Hysteresis is included to filter noise jitter that would otherwise result when v_C and v_{CDLY} cross slowly. As a result, v_C must fall below v_{CDLY} by another 100 mV before CP_{PCH-Det} can trip again after v_{PCH} turns high. A deglitch circuit, shown in Figure 6.16, delays enable signals to hold node v_{O12} down to ground until sufficient current flows through the circuit and, in this way, ensure proper power up. In addition, similar to other comparators in the IC, only one decision from CP_{PCH-Det} matters, which is when v_{PCH} trips high. Class-A output stage MN_2’s sinking current transitions v_{PCH} high quickly, while the other transition is slew-limited by its bias current. However, the system never utilizes the slower slew-limited transition since CP_{PCH-Det} disables right after v_{PCH} turns high. When operational, CP_{PCH-Det} initiates
precharge when $C_{VAR}$ reaches $C_{MAX}$ and disables itself afterwards until the onset of the next cycle.

### 6.2.4 Nanoampere Current Generator

Subthreshold currents generated by the circuit in Figure 6.17 bias the precharge- and harvest-detect comparators $C_{PCH-DET}$ and $C_{HARV-DET}$. The circuit induces gate-source voltage difference between subthreshold-operated NMOS transistors $MN_{P1}-MN_{P2}$ across $R_{PTAT}$ to generate a PTAT current. Similarly, forcing $MN_{P1}$’s gate-source voltage across $R_{CTAT}$ generates a CTAT current through $R_{CTAT}$ and $MP_{C3}$ that, when combined with $MP_{P3}$’s PTAT current, produces a first-order temperature-compensated current $I_{B}$.

The main challenge here is low currents demand high resistances, which is why the circuit uses 50- and 570-MΩ off-chip thick-film resistors for $R_{PTAT}$ and $R_{CTAT}$. Note that other state-of-the-art subthreshold designs require lower resistances by only deriving current from a PTAT voltage, which is in the 50-mV range, rather than a CTAT voltage,
which is around 500 mV, the purpose of which is to temperature-compensate bias currents and avoid higher currents and, therefore, power losses at high temperatures [202]-[203]. As temperature increases, however, leakage currents from diffusion junctions across the IC could overcome the low nanoampere bias currents generated, so, nevertheless, it is important to consider using exclusively PTAT bias currents to overcome leakage currents at high temperatures, even at the expense of greater energy losses.

The nanoampere-current generator powers up with the system and is always operational because, when not biasing a circuit in one phase, it powers another. Therefore, to minimize energy losses, the startup circuit must either shut off completely during normal operation or sink a negligibly low current. With the latter, the small leakage current reverse-biased p+/n-well diode D_{ST} in Figure 6.17 produces biases long-
length diode-connected NMOS transistors \( MN_{S1} - MN_{S2} \) to establish a reference voltage for \( MN_{S3} \). The idea is for \( MN_{S3} \) to prevent \( MN_{P1} - MN_{P2} \) from shutting off by sourcing current into \( C_{C1} \) when \( MN_{C1} \)'s gate voltage attempts to drop, which indicates the generator is liable to enter a zero-current state. In addition, \( C_{C1} \) and \( C_{C2} \) keep noise transients from inadvertently engaging \( MN_{S3} \). Note \( MN_{C1} \)'s gate voltage is sufficiently high during normal operation to keep \( MN_{S3} \) off. Lastly, a monitor circuit like in Figure 6.8 (where italic values represent the nanoampere version) ascertains when the generator is ready to prompts the system to proceed with startup.

### 6.2.5 Power-Up Sequence

The harvesting system synchronizes to the variations in \( C_{VAR} \) by (i) waiting until its nanoampere-current generator is ready and (ii) subsequently discharging and initializing \( C_{VAR} \) to 0 V. The precharger then charges \( C_{VAR} \), irrespective of its value. If \( C_{VAR} \) happens to be decreasing, harvest-detect comparator \( CP_{HARV-DET} \) senses when \( C_{VAR} \) reaches \( C_{MIN} \) when \( i_{HARV} \) is 0 A and prompts the system to cycle through the ensuing reset and precharge phases, even if the first cycle harnesses little to no energy. Conversely, if the system starts when \( C_{VAR} \) is increasing, a reverse harvesting current discharges the battery slightly while causing \( v_C \) to fall below \( V_{BAT} \), which triggers \( CP_{HARV-DET} \) and synchronizes the system into a reset phase. In this way, after one or two irregular cycles, the system starts and synchronizes to \( C_{VAR} \), irrespective of its initial value.

The IC powers up when external digital signal \( v_{PWRU-IC} \) turns high and its rising edge sets latch \( SRST \) in Figure 6.18, consequently, driving transistors \( MN_{DCH1} \) and \( MN_{DCH2} \) to discharge \( C_{VAR} \). The nanoampere bias-current generator then starts up and
Fig. 6.18. Power-up logic (transistor dimensions are in µm).

flags the system it is ready when signal \( v_{\text{RDY-NA}} \) transitions high, which resets the latch and turns off \( MN_{\text{DCH1}} \). In addition, since both signals are high, logic gate \( \text{AND}_{\text{ST}} \)'s output rises, triggering through the rise-edge detector 10-ns pulse signal \( v_{\text{PCH-ST}} \) that forces precharge signal \( v_{\text{PCH}} \) in Figure 6.9 high and, as a result, provoking the start of a precharge phase. Note that during normal operation, since the output of \( \text{AND}_{\text{ST}} \) stays high, \( v_{\text{PCH-ST}} \) is always low and \( v_{\text{PCH}} \) follows the normal precharge signal \( v_{\text{PCH-NM}} \) from the precharge detector.

### 6.2.6 Test Mode, ESD Protection, and Layout

The IC starts up when external digital signal \( v_{\text{PWRU-IC}} \) turns high, but otherwise, the system remains disabled, under test-mode conditions, where each system component, such as comparators \( CP_{\text{VC}} \), \( CP_{\text{SW}} \), \( CP_{\text{PCH-DET}} \), and \( CP_{\text{HARV-DET}} \) and the precharge- and nanoampere-current generators, can be enabled and engaged independently for individual testing. In this way, the user can measure and evaluate the performance of each component independently and the energy consumption of each block is extrapolated by multiplying its measured current draw, battery voltage, and steady-state turn-on time, as
will be presented in the following Chapter 7. Note that in this mode, transistor MN\textsubscript{DCH2} remains off and avoids accidentally shorting node v\textsubscript{C} to ground.

Internal pin-out buffers drive digital signals important to the functionality of the system, such as v\textsubscript{HARV}, v\textsubscript{PCH}, v\textsubscript{GN}, v\textsubscript{GP}, v\textsubscript{HV-END}, v\textsubscript{E-END}, and v\textsubscript{D-END}, off chip to allow debugging and experimental evaluation. What is more, the IC integrates two versions of each detection comparator, \textit{CP\textsubscript{PCH-DET}} and \textit{CP\textsubscript{HARV-DET}}, to accommodate two different vibrations frequencies, 30 Hz and 1.3 kHz (that is, a slow and a fast mode). However, the system was only tested under 30-Hz vibrations because a prototype with higher resonant frequency was not available for measurement.

The circuit in Figure 6.19(a) protects all pins, except the v\textsubscript{SW} node, from electrostatic discharge (ESD) events that occur while handling the IC by sinking to ground, away from the internal components, the large induced currents. For instance, an ESD strike increases the voltage at the bondpad enough to turn on diode D\textsubscript{ESD} (higher than its reverse breakdown voltage), which initially sinks some of the induced current to ground through resistor R\textsubscript{ESD}. Consequently, the voltage drop produced across R\textsubscript{ESD} turns on NPN transistor Q\textsubscript{ESD} to sink the remaining ESD-generated current to ground, at the
bondpad and away from the internal circuits of the IC. In addition, pins that connect
directly to a polysilicon gate are further protected with the local clamp in Figure 6.19(b)
that, when placed close to the target gate, limits the induced voltage by turning on NPN
$Q_{\text{CLAMP}}$ in reverse-active mode, which features a lower breakdown voltage than when in
forward-active mode [204]. Resistor $R_{\text{CLAMP}}$ limits current flowing to the gate.

Guard rings surround diffusions that connect to pins off chip to protect the IC
from transient-induced latchup. For instance, node $v_{\text{SW}}$ connects to the drains of power
switches $M_{\text{PE}}$ and $M_{\text{ND}}$ and swings from below ground to above $V_{\text{BAT}}$ during precharge,
injecting minority carriers into the n-well of $M_{\text{PE}}$ and the IC substrate through $M_{\text{ND}}$. For
this reason, a hole-blocking guard ring surrounds $M_{\text{PE}}$, where a deep-n+ diffusion
surrounds the transistor perimeter while a n-type buried layer (NBL) covers its bottom,
both tied to the n-well supply, which is $V_{\text{BAT}}$. The NBL diffusion blocks holes from
travelling downward to the substrate while the surrounding deep-n+ diffusion prevents
holes from travelling laterally.

Similarly, the same diffusion layers, deep-n+ and NBL, make up an electron-
collecting guard ring around $M_{\text{ND}}$ that recombines electrons flowing into the p-type
substrate and travelling laterally away from the transistor. Both diffusions tie to ground to
avoid unwanted current flow, and therefore energy draw, from $V_{\text{BAT}}$, although connecting
to a higher voltage would make the guard ring depletion region deeper and more
effective. Nevertheless, both guard rings prevent injected minority carriers from
travelling through the p-type substrate and n-well and potentially latching up and
adversely affecting other devices in the system that share the same substrate and well.
In conclusion, the presented energy-harvesting IC was implemented in a 0.7-µm BiCMOS process, as seen in the floorplan in Figure 6.20 and layout diagram in Figure 6.21. The IC features 32 pins, most of which, however, are only used for experimental testing and measurement (e.g., buffered digital signals and test-mode setting inputs). The
entire system, including redundant comparators, test-mode logic, pin-out digital buffers, testpads, ESD protection circuits, and bondpads (80 x 112 µm² each), was laid out inside 1 x 1 mm² of silicon area.

6.3 Summary

An electrostatic energy-harvesting system IC was presented that precharges and holds the voltage across variable capacitor C_{VAR}, while ambient vibrations induce and generate harvesting current i_{HARV} into the battery as capacitance decreases. The energy harvested must exceed all losses in the system, including not only the battery investment to precharge C_{VAR}, but also the energy the system utilizes to coordinate and synchronize each phase in the energy-harvesting cycle. This includes detecting C_{MAX} during reset (precharge detection), precharging C_{VAR} to V_{BAT} (precharger control), connecting C_{VAR} to the battery during the harvesting phase (harvesting switch MP_H), and sensing C_{MIN} to disconnect C_{VAR} from everything during reset (harvest detection). To minimize energy consumption, the IC only enables and powers each component when necessary and shutting each off when unused. System blocks that engage for longer fractions of the vibration cycle, such as each of the detection subsystems, operate with subthreshold currents. The proposed energy-harvesting IC, as a result, is capable of synchronizing to C_{VAR} while minimizing the negative impact it has on energy gain.
CHAPTER 7

ENERGY HARVESTER IC PROTOTYPE RESULTS

The prototyped energy-harvesting system was integrated into the 0.7-μm BiCMOS 1 x 1 mm\(^2\) silicon die in Figure 7.1, packaged in a 32-pin plastic 7 x 7 mm\(^2\) quad-flat package (PQFP), and tested on the printed-circuit board (PCB) in Figure 7.2. The IC includes the entire system, as presented in Figure 6.2, except L, C\(_{VAR}\), and for testing purposes, precharge reference V\(_{REF}\) and bias, delay, and sense resistors R\(_{PTAT}\), R\(_{CTAT}\), R\(_{DLYA}\), and R\(_{HARV}\). The IC also incorporates test-only circuits such as pin-out digital buffers, extra test-mode logic, and redundant comparators. On the PCB board, thick-film resistors compose R\(_{PTAT}\), R\(_{CTAT}\), and R\(_{DLYA}\) since they feature significant resistance values greater than 10 MΩ, while a 100-kΩ surface-mount 0805 resistor makes up R\(_{HARV}\) (corresponding to a 2.00 x 1.25 mm\(^2\) area). These resistors were untrimmed, and their values were unchanged in all samples measured. No action was taken to

Figure 7.1. Die photograph of the 1 x 1 mm\(^2\) energy-harvesting IC.
improve accuracy because, in simulations, a 20% variation in $R_{DLY}$ corresponded to only 1.7% variation in detecting $C_{MAX}$ since $C_{PCH-DET}$ includes sufficient input-referred offset and voltage-gain margin to accommodate the 20% voltage variation that results between its input terminals $v_C$ and $v_{CDLY}$. Furthermore, 20% variations in $R_{PTAT}$ and $R_{CTAT}$ had negligible impact on the total energy harvested, at most 4.7% change.

Other external PCB components include the 2 x 2 x 1 mm$^3$ 10-µH Coilcraft EPL2010 precharge inductor, which introduced a maximum equivalent series resistance
(ESR) of 1 Ω. Also, while on-chip pin-out digital buffers drive internal digital signals \(v_{\text{HARV}}, v_{\text{PCH}}, v_{\text{GN}}, v_{\text{GP}}, v_{\text{HV-END}}, v_{\text{E-END}}, \) and \(v_{\text{D-END}}\) off chip, analog signals important for measurement and debugging, such as \(v_C\) and \(v_{\text{SW}}\), are buffered with external 250-MHz low-offset (< 8 mV) OPA2357 amplifiers in unity-gain configuration, as shown in Figure 7.3. The PCB test amplifiers are powered by a separate supply that has sufficient headroom above \(V_{\text{BAT}}\) and below ground (since \(v_{\text{SW}}\) swings below ground), while internal buffers are powered by an additional supply at the same voltage level as the internal logic (which is \(V_{\text{BAT}}\)). In this way, oscilloscope probe measurements do not affect any of the IC nodes with its equivalent 10-MΩ and 10-pF load.

The dc bias currents flowing into each of the system’s components were measured with Keithley’s Model 6485 “picoammeter”, which is capable of accurately discerning currents from as low as 20 fA to as high as 20 mA, with errors of less than 0.4 %. Its wide current range not only allows measurement of the extremely diminutive nA-range currents that the nanoampere bias-current generator and detection comparators \(CP_{\text{PCH-DET}}\) and \(CP_{\text{HARV-DET}}\) require, but also the more usual µA-range currents that the precharge

![Figure 7.4. Vibration-driven variable capacitor prototype.](image)
control circuits use (e.g., CP\textsubscript{SW}, CP\textsubscript{VC}, and precharge bias-current generator).

7.1 \(C_{\text{VAR}}\) Prototype

The prototyped 10.16 \(\times\) 20.32 cm\(^2\) (4 \(\times\) 8 in\(^2\)) and 0.125 kg \(C_{\text{VAR}}\) in Figure 7.4 emulates a MEMS counterpart and permits testing the energy-harvesting system. The vibration-driven variable capacitor prototype features a top plate of two 10.16 \(\times\) 20.32 \(\times\) 0.013 cm\(^3\) (4 \(\times\) 8 \(\times\) 0.005 in\(^3\)) 1095-spring steel sheets and a 10.16 \(\times\) 20.32 \(\times\) 0.046 cm\(^3\) (4 \(\times\) 8 \(\times\) 0.018 in\(^3\)) steel bottom plate. Three non-conducting nylon screws with separating 0.1 cm-thick nylon washers connect the plates across their centerline axis. Before testing the IC, the inverting op amp in Figure 7.5(a) measures, via output \(v_{\text{OUT}}\), \(C_{\text{VAR}}\) as it shakes by amplifying high-frequency input \(v_{\text{IN}}\) by \(C_{\text{VAR}}/C_{\text{REF}}\) [205], as the results in Figure 7.5(b) show. Accordingly, the gain across the circuit is a direct measure of \(C_{\text{VAR}}\), when using a well-characterized reference capacitor \(C_{\text{REF}}\), which is 98 pF plus 14 pF from board parasitics. Ultimately, measurements show \(C_{\text{VAR}}\) resonates at 30 Hz and varies from 165.8 to 967.7 pF when shaken at the middle screw by a Brüel & Kjær 4810

![Diagram](image)

Figure 7.5. (a) Capacitance-sensing circuit and (b) corresponding measurement results.
vibration source with an estimated acceleration of approximately 70 m/s² (based on the shaker specifications and not directly measured). Note however that typical environments feature accelerations below 12 m/s² [95].

7.2 Harvest and Reset Phases

Since the Keithley low-current ammeter that measures dc bias currents does not evaluate transient signals, the 100 V/V LTC1100 instrumentation amplifier IA_H in Figure 7.6(a) augments the voltage across R_HARV (100 kΩ), which directly correlates to i_HARV, by its gain at output v_I_HARV. The amplifier features less than 0.075% of gain error and 10 µV of input-referred voltage offset. The current-sourcing circuit in Figure 7.6(b), which was built in a separate PCB board along with the capacitance-sensing circuit in Figure 7.5(a), verifies the accuracy of the setup by generating current i SOURCE that is then measured by the i_HARV-sensing circuit. Measuring output v_I_HARV with an oscilloscope at 1 V/div resolution yields a consistent average offset error of 8.29 nA across the current range of 0-600 nA, when compared to the Keithley ammeter measurement (with less than 0.4% error). By calibrating the oscilloscope, the average error was reduced to 0.29 nA across the entire range, as the results in Figure 7.7 show, where i_CAL and i_UNCAL represent the

![Diagram](image)

Figure 7.6. (a) i_HARV-sensing circuit and (b) current source used to measure its accuracy.
calibrated and uncalibrated measurements, respectively. To corroborate the measurement further, \( v_{i_{HARV}} \) was also measured with an accurate dc multimeter that yielded an error of 0.34 nA across the measurement range. One problem with the setup, however, is that \( R_{HARV} \) connects directly to \( C_{VAR} \), introducing about 25 pF of measured parasitic capacitance. Future implementations, instead, connect \( R_{HARV} \) and the current-sensing instrumentation amplifier to the battery side of harvesting switch \( MP_H \) to avoid the additional energy investment the parasitic capacitance demands.

As the \( C_{VAR} \) prototype vibrates at 30 Hz, the IC synchronizes to it during each phase: harvest, reset, and precharge. As Figure 7.8 shows, during harvest, vibrations, through \( C_{VAR} \), generate up to 500 nA when using a battery at 3.5 V. As each harvesting phase ends, \( i_{HARV} \) reduces to 0 A and reset follows with \( C_{VAR} \)'s voltage \( v_C \) gradually dropping from its harvesting state of 3.5 V to a minimum, as Figure 7.8(a) confirms. Harvesting control signal and \( MP_H \) gate voltage \( v_{HARV} \) in Figure 7.8(b) transitions

Figure 7.7. Calibrated and uncalibrated current measurements using the \( i_{HARV} \)-sensing circuit.
accordingly, with a low state engaging $MP_H$ to connect $C_{VAR}$ to the battery and a high state prompting the system to enter reset.

Harvesting current $i_{HARV}$ represents, when integrated over time as it flows into the 3.5 V battery, an average gain of 11.11 nJ/cycle. The harvesting detector introduces a brief delay at the end of the harvesting phase that allows $C_{VAR}$ to increase slightly while still connected to the battery, drawing a reverse current that discharges the battery by

Figure 7.8. Experimental measurements showing (a) variable capacitor voltage $v_C$, harvesting current $i_{HARV}$, extrapolated energy gain $E_{HARV}$, and (b) harvesting control signal $v_{HARV}$ during five vibration cycles.
342.64 pJ/cycle. The detector, which derives power from the 3.5 V battery and is active through the harvesting phase, which is about 20.0 ms/cycle on average, consumes a measured quiescent current $I_Q$ of 2.30–3.32 nA, resulting in an average dissipation of 190.84 pJ/cycle. Similarly, the precharge detector draws a measured $I_Q$ of 1.13–3.34 nA for the duration of the reset phase, or approximately 13.3 ms/cycle on average, resulting in an average dissipation of roughly 94.53 pJ/cycle. The vibration period is on average 33.3 ms and its corresponding frequency is 30.0 Hz. The nanoampere bias-current generator, which biases both detection blocks and remains operational through the entire period, sinks 3.22–3.72 nA from the 3.5 V supply, dissipating an average of 400.85 pJ/cycle. As Table 7.1 summarizes, the battery gains 10.58 nJ/cycle during harvesting, loses 94.53 pJ/cycle in reset, and loses another 400.85 pJ/cycle to the nanoampere current generator.

Table 7.1. Measured energy consumed and gained by the prototyped harvester IC.

<table>
<thead>
<tr>
<th>Phase</th>
<th>Measured Energy (nJ/cycle)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$V_{BAT} = 2.7$ V</td>
</tr>
<tr>
<td>Harvest Phase</td>
<td></td>
</tr>
<tr>
<td>Harvested Energy</td>
<td>+7.047</td>
</tr>
<tr>
<td>Reverse Energy</td>
<td>-0.233</td>
</tr>
<tr>
<td>Control Circuits</td>
<td>-0.148</td>
</tr>
<tr>
<td>Reset Phase</td>
<td></td>
</tr>
<tr>
<td>Control Circuits</td>
<td>-0.069</td>
</tr>
<tr>
<td>Precharge Phase</td>
<td></td>
</tr>
<tr>
<td>Invested Energy</td>
<td>-5.629</td>
</tr>
<tr>
<td>Returned Energy</td>
<td>+0.353</td>
</tr>
<tr>
<td>Sense Resistor</td>
<td>+0.304</td>
</tr>
<tr>
<td>Control Circuits</td>
<td>-0.048</td>
</tr>
<tr>
<td>nA Bias-Current Generator</td>
<td>-0.304</td>
</tr>
<tr>
<td>Net Energy Gain (per cycle)</td>
<td>+1.273 nJ</td>
</tr>
<tr>
<td>Power Gain (at 30 Hz)</td>
<td>+38.19 nW</td>
</tr>
</tbody>
</table>
7.3 Precharge Phase

Variable capacitor $C_{\text{VAR}}$ charges from close to 0 V to its 3.5 V target between every reset and harvesting phase, as the results in Figure 7.8(a) demonstrate. The variable capacitor’s voltage $v_C$ in Figure 7.9(a) first rises to roughly 3 V when the system energizes $L$ and $C_{\text{VAR}}$ from the battery, as energizing switch $MP_E$’s low gate voltage $v_{GP}$ from Figure 7.9(b) induces switching voltage $v_{SW}$ in Figure 7.9(a) to remain high.

Figure 7.9. Precharge waveforms showing (a) variable capacitor voltage $v_C$ and switch-node voltage $v_{SW}$ with corresponding (b) energize and de-energize gate-control signals $v_{GP}$ and $v_{GN}$.
Capacitor voltage $v_C$ then reaches 3.5 V when the system de-energizes L into $C_{VAR}$ as de-energizing switch MN$_D$’s high gate voltage $v_{GN}$ forces $v_{SW}$ to stay low. The system introduces an average dead time between MP$_E$ and MN$_D$ conduction events of 1.86 ns.

Reference voltage $V_{REF}$, which sets $v_C$’s energizing target and, therefore, the energizing time, was manually adjusted to ensure $v_C$ reached $V_{BAT}$ at the end of precharge. As before, although $V_{REF}$ should be in theory $0.5V_{BAT}$, it was higher because
the IC required more investment energy $E_{INV}$ to compensate for the power lost in the circuit. As a result, because process variations across dice introduced parameter variations in the chips tested, $V_{REF}$ varied between 2.5 and 2.8 V across prototyped ICs when tested at 3.5 V. In the end, the system energized $L$ and $C_{VAR}$ for about 155.44 ns on average, producing a peak inductor current of 23.34 mA that subsequently dropped to 0 A, while de-energized, in 79.23 ns. The average end-of-precharge voltage is 3.79 V. Note switching node $v_{SW}$ dropped to about $-783$ mV and increased to 0 V during the de-energizing step. Also notice remnant energy in $L$ and adjacent parasitic capacitors produced oscillations in $v_{SW}$, which the system eventually dampened.

Precharge-detect comparator $CP_{PCH-DET}$’s output voltage $v_{PCH}$ in Figure 7.10(a) rises when $C_{VAR}$ reaches $C_{MAX}$ to prompt the system to power the precharge bias-current generator, which becomes fully functional after approximately 262.83 ns. Once biased, the logic circuits initiate the energize/de-energize sequence that charges $C_{VAR}$, via gate-control signals $v_{GP}$ and $v_{GN}$ in Figure 7.10(b), and powers both precharge comparators, enabling $CP_{SW}$ only after the de-energize step begins. Both comparators and their current generator remain biased through the end of the precharge phase, marked by the fall of $v_{HARV}$. In the end, the current generator remains operational for about 499.63 ns, drawing 12.01–12.82 µA from the 3.5 V battery and using approximately 21.62 pJ/cycle.

Energizing comparator $CP_{VC}$’s output $v_{E-END}$ remains low until $v_{C}$ reaches $V_{REF}$ and de-energizing comparator $CP_{SW}$’s output $v_{D-END}$ stays high until the end of the de-energizing step, as shown in Figure 7.11(a). $CP_{VC}$ draws 22.13–24.91 µA for about 141.29 ns while $v_{E-END}$ is low and 29.33–34.62 µA for an additional 95.51 ns until precharge ends. Similarly, $CP_{SW}$ draws about 20.12–28.85 µA while it holds $v_{D-END}$ high
for 205.57 ns. Built-in input offset $V_{OS}$, which is 153.43 mV on average, causes $CP_{SW}$ to trip low slightly before the de-energizing step ends, for about 31.23 ns. During this time, $CP_{SW}$ draws 16.74–23.78 µA. In all, $CP_{VC}$, $CP_{SW}$, and their bias generator together demand 63.65 pJ/cycle, of which $CP_{VC}$ dissipates 22.16 pJ/cycle and $CP_{SW}$ 19.88 pJ/cycle.

Figure 7.11. (a) $CP_{VC}$ and $CP_{SW}$’s outputs, $v_{E-END}$ and $v_{D-END}$, which mark the end of the energize and de-energize steps, respectively, and (b) precharge current $i_{PCH}$ with extrapolated investment energy $E_{INV}$.
The average energy the battery invested into the system to precharge $C_{VAR}$, $E_{INV}$, in Figure 7.11(b), was 9.03 nJ/cycle, as measured from the battery current drawn through a series 10 Ω sense resistor. Recall $E_{INV}$ includes the losses the logic switches and gate drivers in the IC incur during precharge. For a 3.5 V battery, $i_L$ increased on average to 23.34 mA, overcharging $C_{VAR}$ to 3.79 V. The additional 290 mV in $C_{VAR}$ caused the system to return the overinvested energy to the battery at the beginning of every harvesting phase, which is why the battery receives back an average of 662.02 pJ/cycle.

<table>
<thead>
<tr>
<th>Precharger</th>
<th>CPVC</th>
<th>I_O, High Output: 16.9-24.2 μA</th>
<th>20.1-28.9 μA</th>
<th>22.2-32.0 μA</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>I_O, Low Output: 13.8-19.9 μA</td>
<td>16.7-23.8 μA</td>
<td>18.7-25.9 μA</td>
<td></td>
</tr>
<tr>
<td></td>
<td>$t_{ON, AVG} (High/Low)$: 199.1/42.6 ns</td>
<td>205.6/31.2 ns</td>
<td>185.1/25.9 ns</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Avg. $</td>
<td>V_{OS}</td>
<td>$: 54.7 mV</td>
<td>57.0 mV</td>
</tr>
<tr>
<td></td>
<td>$t_{dqy, AVG}$: 7.89 ns</td>
<td>5.99 ns</td>
<td>5.57 ns</td>
<td></td>
</tr>
<tr>
<td>Precharge Control</td>
<td>CPsw</td>
<td>I_O, High Output: 16.9-24.2 μA</td>
<td>20.1-28.9 μA</td>
<td>22.2-32.0 μA</td>
</tr>
<tr>
<td></td>
<td>I_O, Low Output: 13.8-19.9 μA</td>
<td>16.7-23.8 μA</td>
<td>18.7-25.9 μA</td>
<td></td>
</tr>
<tr>
<td></td>
<td>$t_{ON, AVG} (High/Low)$: 199.1/42.6 ns</td>
<td>205.6/31.2 ns</td>
<td>185.1/25.9 ns</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Avg. $</td>
<td>V_{OS}</td>
<td>$: 54.7 mV</td>
<td>57.0 mV</td>
</tr>
<tr>
<td></td>
<td>$t_{dqy, AVG}$: 7.89 ns</td>
<td>5.99 ns</td>
<td>5.57 ns</td>
<td></td>
</tr>
</tbody>
</table>

The average energy the battery invested into the system to precharge $C_{VAR}$, $E_{INV}$, in Figure 7.11(b), was 9.03 nJ/cycle, as measured from the battery current drawn through a series 10 Ω sense resistor. Recall $E_{INV}$ includes the losses the logic switches and gate drivers in the IC incur during precharge. For a 3.5 V battery, $i_L$ increased on average to 23.34 mA, overcharging $C_{VAR}$ to 3.79 V. The additional 290 mV in $C_{VAR}$ caused the system to return the overinvested energy to the battery at the beginning of every harvesting phase, which is why the battery receives back an average of 662.02 pJ/cycle.
In other words, when harvesting switch $M_{PH}$ first closes, $i_{PCH}$ peaks at $-3.6 \text{ mA}$ and $E_{INV}$ drops to roughly $8.37 \text{ nJ/cycle}$. In all, the system invested $7.88 \text{ nJ/cycle}$. Note the budget in Table 7.1 adds back the $485.04 \text{ pJ/cycle}$ the $10 \Omega$ sense resistor dissipated because the only reason the system required this energy in the first place was to test it. If the investment efficiency of the precharger is defined as the ratio of ideal to actual energy investment, then the results show that at $3.5 \text{ V}$ it is $75.22 \%$ (as the ideal investment to $967.7 \text{ pF}$ is $5.927 \text{ nJ}$). Table 7.2 summarizes the IC’s experimental performance.

### 7.4 Energy Gain

The total energy the system drew from vibrations in $C_{VAR}$ exceeded all losses to produce a gain of $2.140 \text{ nJ/cycle}$ for a $3.5 \text{ V}$ battery, which is equivalent to $64.20 \text{ nW}$ at $30 \text{ Hz}$. The system also produced gains of $1.273$ and $2.869 \text{ nJ/cycle}$ at $2.7$ and $4.2 \text{ V}$, which represents the operating range of typical Li Ions, for $38.19 \text{ nW}$ and $86.07 \text{ nW}$, respectively. The IC charges $1-\mu\text{F}$ capacitor $C_{BAT}$, which emulates a microscale battery, from $3.5 \text{ V}$ to $3.81 \text{ V}$ (that is, $1.13 \mu\text{J}$) in $35.16 \text{ s}$ when setting $V_{REF}$ to $2.8 \text{ V}$, as pictured in Figure 7.12. For this measurement, off-chip amplifier TL074 buffers $C_{BAT}$’s voltage $v_{CBAT}$ from the oscilloscope probe. Note that a real battery, such as a $1\text{-mAh}$ thin-film Li Ion, has substantially higher capacity than $1 \mu\text{F}$ and charging it at the slow rates that the harvester IC can would take considerably longer (up to several days).

As $C_{BAT}$’s voltage increases, $C_{VAR}$ should precharge to an increasingly higher level, demanding $V_{REF}$ to increase accordingly. Because $V_{REF}$ is fixed, however, the system eventually is unable to invest sufficient energy into $C_{VAR}$ to avoid $v_{C-V_{BAT}}$ mismatch voltage losses across $M_{PH}$ from increasing to the point a gain is no longer possible. In addition, when $C_{VAR}$ is severely undercharged, although it finishes
precharging through $MP_H$, $CP_{HARV-DET}$ can prematurely trip before $v_C$ reaches $V_{BAT}$, causing the system to undergo additional reset and precharge phases that augment energy losses. In other words, the harvest detection block incorrectly senses $C_{VAR}$ reached $C_{MIN}$, but since $C_{VAR}$ is actually decreasing, $CP_{PCH-DET}$ immediately forces an additional precharge phase. The result is the system stops charging $C_{BAT}$ at 3.81 V. Fixing the harvest-detection subsystem to avoid premature triggering and including a feedback loop to dynamically tune $V_{REF}$ to ensure $v_C$ reaches $V_{BAT}$ during precharge would eliminate the problem, except the losses in the same must be low enough for the system to continue generating a net positive gain.

### 7.5 Discussion

The fact the prototype generated and channeled 2.140 nJ/cycle (64.20 nW at 30 Hz) into a 3.5 V battery means the system can replenish some of the energy a wireless microsensor consumes. The generated power may seem low for practical applications but duty-cycling the sensor to operate a fraction of the time viably enables the system to
accumulate sufficient energy in the battery to supply the power needed when demanded. That is, the on-board battery powers the sensor’s high-power tasks, such as wireless transmission, only after the battery amasses enough energy from the harvester. Consider, for instance, that 10 ms of wireless telemetry at 5 mW (5 ms for transmission and 5 ms for reception) and sensing at 10 µW for 1 ms requires 50.01 µJ [36]. According to the energy harvested from a 3.5 V battery, the harvester can replenish the total energy used in 778.97 s (12.98 min.). Similarly, with $V_{\text{BAT}}$ at 2.7 V and 4.2 V, the harvester replenishes the total energy used in 1309.51 s (21.83 min.) and 581.04 s (9.68 min.), as Table 7.3 summarizes. In other words, allowing the wireless microsensor to sense and transmit once every 13 minutes, when the battery voltage is at 3.5 V, enables the prototype to harvest from the environment all the energy the system requires, extending its operational life almost indefinitely, barring the wear-and-tear effects of the components.

The fundamental advantage of constraining $C_{\text{VAR}}$’s voltage instead of its charge is sub-5 V operation because the extreme voltages that restraining charge would otherwise produce require higher voltage transistors, which only lower volume and higher cost semiconductor technologies offer. Another benefit is using the already existing battery-

Table 7.3. Estimates of microsensor consumption and energy-harvester IC requirements.

<table>
<thead>
<tr>
<th>Load Duty-Cycle Profile</th>
<th>System Tasks</th>
<th>Duration</th>
<th>Power</th>
<th>Energy</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Transmission</td>
<td>5.00 ms</td>
<td>-5.00 mW</td>
<td>-25.00 µJ</td>
</tr>
<tr>
<td></td>
<td>Reception</td>
<td>5.00 ms</td>
<td>-5.00 mW</td>
<td>-25.00 µJ</td>
</tr>
<tr>
<td></td>
<td>Sensing</td>
<td>1.00 ms</td>
<td>-10.00 µW</td>
<td>-0.01 µJ</td>
</tr>
<tr>
<td>Total Consumption</td>
<td></td>
<td></td>
<td></td>
<td>-50.01 µJ</td>
</tr>
<tr>
<td>Harvester Requirement at 2.7 V:</td>
<td>&gt; 1309.51 s</td>
<td>+38.19 nW</td>
<td>&gt; +50.01 µJ</td>
<td></td>
</tr>
<tr>
<td>Harvester Requirement at 3.5 V:</td>
<td>&gt; 778.97 s</td>
<td>+64.20 nW</td>
<td>&gt; +50.01 µJ</td>
<td></td>
</tr>
<tr>
<td>Harvester Requirement at 4.2 V:</td>
<td>&gt; 581.04 s</td>
<td>+86.07 nW</td>
<td>&gt; +50.01 µJ</td>
<td></td>
</tr>
</tbody>
</table>
to-be-charged to constrain voltage. These functional gains, however, result at the expense of lower energy because energy is proportional to voltage. Nevertheless, drawing low power over time can ultimately harvest vast amounts of energy, which duty-cycled microsensors can viably manage and endure.

The prototype suffers from a few disadvantages that an otherwise improved design could relinquish. To start, as previously mentioned, a low-bandwidth feedback loop should dynamically adjust $V_{\text{REF}}$ to ensure the system charges $C_{\text{VAR}}$ to $V_{\text{BAT}}$, even as $V_{\text{BAT}}$ changes and/or other system conditions vary. In addition, the presented results do not account the losses associated with an on-chip fixed $V_{\text{REF}}$, even though these should be low if the reference operates in subthreshold or only engages during the precharge phase. The designer should also optimize the speed and losses of the IC to operate at the known vibration frequency and $C_{\text{VAR}}$’s range. In the presented case, $C_{\text{VAR}}$’s resonance frequency and capacitance range were unavailable during the design phase so optimizing the precharger’s switching-conduction tradeoff losses was difficult. Additionally, operating the detection circuits for only a fraction of each half cycle would reduce losses, even though, minimizing the operation time of the detection circuits presents the challenge in that creating sufficiently long turn-on delays, since each vibration period is relatively long at 33.3 ms, requires substantial capacitance and therefore considerable energy. Finally, including battery-protection features by monitoring $V_{\text{BAT}}$ every several vibration cycles would complete the system at a small incremental energy expense.

Note that the prototype outputs usable power to the microsystem even when using a non-optimal energy-conversion device. More efficient transduction schemes maximize mechanical-electrical energy conversion by minimizing mechanical losses like air
friction. From an electrical standpoint, a transducer optimized for voltage-constrained harvesting seeks to maximize its capacitance difference $\Delta C_{\text{VAR}}$, rather than increase its maximum-to-minimum ratio $C_{\text{MAX}}/C_{\text{MIN}}$, which benefits charge-constrained systems. When integrated, a MEMS device would have to also manage low pull-in voltages, stiction, and relatively large areas to realize high capacitances, so an optimized solution will more than likely exhibit both a smaller $\Delta C_{\text{VAR}}$ and a lower $C_{\text{MAX}}$ (e.g., $\Delta C_{\text{VAR}} \approx 500–100$ pF) and, because of the lighter mass, resonate at higher frequency. These characteristics partially compensate one another because, while a smaller $\Delta C_{\text{VAR}}$ reduces the energy harvested, a lower $C_{\text{MAX}}$ requires less investment energy and a shorter period decreases the time, and therefore energy, detection blocks operate in each cycle. Shorter vibration periods, however, require faster comparators and proper adjustment of the precharge detection RC delay circuit. Nevertheless, the aim and significance of the presented prototype is to convert as much energy as the transducer avails, irrespective of the quality and efficiency of the transducer, research for which others better trained in the art currently conduct.

The design used the 5-V n- and p-type MOSFETs and 8-V NPN BJTs that the high-volume BiCMOS process availed. Only two instances in the entire system exploit the bipolar features offered: (1) the p-tank that embeds $C_{\text{P}}$’s n-type input pair in Figure 6.6 and (2) the vertical NPN BJTs from the precharge bias-current generator in Figure 6.7. Although these choices reduce noise sensitivity, improve bias accuracy, and use smaller transistors, an all-CMOS design that allows the NMOS pair to lie in the p-type substrate and employs large subthreshold MOSFETs in place of BJTs (as the nanoampere bias-current generator in Figure 6.17 does) would work. The CMOS solution could also
integrate on chip the external voltage reference and off-chip resistors the prototyped system used for testing flexibility and proof-of-concept purposes.

7.6 Summary

The presented IC harvested 1.273, 2.140, and 2.869 nJ/cycle from vibrations at 30 Hz, generating 38.19, 64.20, and 86.07 nW, and used the energy to charge a battery at 2.7, 3.5, and 4.2 V and charge a 1-µF battery-emulating capacitor from 3.5 to 3.81 V in 35 s. The system did this by efficiently sensing and synchronizing a variable capacitor’s state as it cycled from \( C_{\text{MAX}} \) and \( C_{\text{MIN}} \) to (i) precharge it at \( C_{\text{MAX}} \), (ii) harvest while it decreases to \( C_{\text{MIN}} \), and (iii) reset automatically as it increases back to \( C_{\text{MAX}} \). Producing a net energy gain, however, ultimately translates to reducing losses, which is why the system time-managed and biased its circuits to operate only when needed and with just enough energy, where some components operate deep in subthreshold. Although dynamically adjusting the precharge target voltage could further reduce losses and increase output power, the system nonetheless produced a net gain that could viably extend the life of a wireless microsensor indefinitely.
CHAPTER 8
SELF-TUNING HARVESTER IC AND PROTOTYPE RESULTS

Vibration-driven variable capacitor $C_{\text{VAR}}$ requires charge to establish the electrostatic force against which vibrations will work to harvest kinetic energy, which is why the battery must invest energy $E_{\text{INV}}$ to precharge $C_{\text{VAR}}$ to battery voltage $V_{\text{BAT}}$ at $C_{\text{MAX}}$. To ensure the system invests sufficient $E_{\text{INV}}$ to raise $v_c$ to $V_{\text{BAT}}$ during precharge, $V_{\text{BAT}}$ should ideally energize $L$ and $C_{\text{VAR}}$ until $v_c$ reaches $0.5V_{\text{BAT}}$. In practice, however, losses increase the energy needed so $v_c$ must therefore rise to a higher voltage that reference voltage $v_{\text{REF}}$ sets when comparator $\text{CP}_{\text{VC}}$ in Figure 8.1 trips. As $V_{\text{BAT}}$ changes, so does the investment requirement $E_{\text{INV}}$, and $v_{\text{REF}}$ should, as a result, track $V_{\text{BAT}}$ to avoid under- or overcharging $C_{\text{VAR}}$ during precharge, which would otherwise impress a higher voltage across $\text{MP}_H$ at the beginning of the harvesting phase and result in significant

Figure 8.1. Self-tuning precharger circuit (transistor dimensions are in $\mu$m).
energy losses. To that end, the proposed self-tuning harvester IC utilizes the self-adjusting precharger in Figure 8.1 to regulate the $v_C$’s final precharge voltage to $V_{BAT}$ and, in this way, adapt to changing conditions and the battery’s state of charge.

8.1 System

The self-tuning harvester regulates how much energy $V_{BAT}$ invests in $L$ and $C_{VAR}$ by tuning on a cycle-by-cycle basis the precharger’s energizing time $t_E$. After each precharge phase, comparator $CP_{REF}$, shown in Figures 8.1 and 8.2, compares $v_C$ to $V_{BAT}$ to determine whether $C_{VAR}$ under- or overcharged. If $C_{VAR}$ is overcharged above $V_{BAT}$, $CP_{REF}$ decreases $v_{REF}$ to reduce $t_E$ and $E_{INV}$ for the subsequent vibration cycle. Conversely, $v_{REF}$ increases if the precharger undercharges $C_{VAR}$ below $V_{BAT}$. The system accordingly tunes $t_E$ by increasing or decreasing $v_{REF}$ after each precharge in preparation for the next cycle. In steady state, therefore, $C_{VAR}$ charges to $V_{BAT}$ accurately, which minimizes conduction losses across $MP_H$.

Figure 8.2. Tuning reference circuit (transistor dimensions are in $\mu m$).
Comparator CP\textsubscript{REF} in Figure 8.2 compares $V_{\text{BAT}}$ and $v_{C}$ only after each precharge phase, shutting off immediately afterwards to conserve energy. Current source $I_{\text{CH}}$ and sink $I_{\text{DCH}}$ pump or remove charge $\Delta Q_{\text{REF}}$ from on-chip reference capacitor $C_{\text{REF}}$ to increase or decrease $v_{\text{REF}}$ by fixed amount $\Delta v_{\text{REF}}$

$$\Delta v_{\text{REF}} = \pm \frac{\Delta Q_{\text{REF}}}{C_{\text{REF}}}.$$  \hspace{1cm} (8.1)

In steady state, $v_{\text{REF}}$ toggles between its two most optimal values for the current $V_{\text{BAT}}$, changing in $\Delta v_{\text{REF}}$ steps to correspondingly adjust the precharger’s energizing time of the next cycle. When the system initializes, however, $v_{\text{REF}}$ rises from ground one $\Delta v_{\text{REF}}$ at a time so the harvester is unable to yield energy until $v_{\text{REF}}$ is within a margin of its optimal state.

The system regulates $v_{C}$’s final precharge voltage by tuning $t_{E}$ with a feedback loop in discrete time. In other words, it operates only during a small fraction of the vibration cycle to generate $v_{\text{REF}}$ for the next cycle by adding or subtracting step $\Delta v_{\text{REF}}$ to the current state. Capacitor $C_{\text{REF}}$ then holds that state for the remainder of the vibration period, until the next cycle, which is when the new $v_{\text{REF}}$ value determines the precharge energizing time. In this way, the loop dissipates power only for a small portion of the period. Including so much time for signals to settle introduces a dominant pole to the loop that decreases the open-loop gain to unity at a frequency that is considerably lower than any other pole in the loop, which is why the circuit is stable. Note the feedback loop breaks with a fixed reference because $v_{C}$ resets and charges to a fixed preset value every cycle.
8.2 Integrated Circuit

The self-tuning energy-harvesting system integrates all blocks, including phase detection and control circuits, into a single silicon IC, with the exception of L, C\text{VAR}, bias current-setting resistors, and R\text{HARV}, which are off chip for experimental flexibility, as Figure 8.3 shows. In addition, precharge switches MP\text{E} and MN\text{D} and their respective gate drivers were resized, as Figure 8.4 illustrates. Comparator CP\text{REF}, which is at the core of

Figure 8.3. Prototype self-tuning energy harvester IC (transistor dimensions are in \(\mu\text{m}\)).
the self-tuning loop, monitors $v_C$ with preamplifier $AMP_{PRE}$ and drives the programmable reference block with latch comparator $CP_{LATCH}$, as shown in Figure 8.5(a). Based on $CP_{REF}$’s output, logic engages $MP_{CH}$ or $MN_{DCH}$ to charge or discharge $C_{REF}$. The designed delay that the rising edge-detector in Figure 8.5(b) sets determines the duration of each charge or discharge. After $v_{REF}$ settles to its new state, the delayed signal drops and the falling-edge detector causes $v_{HARV-START}$ to trip high (Figure 8.2), engaging switch $MP_H$ to close and start the harvesting phase.

### 8.2.1 Charge Pump

While $CP_{REF}$’s outputs $v_{O^+}$ or $v_{O^-}$ determine whether to charge or discharge poly-poly capacitor $C_{REF}$ with currents $I_{CH}$ or $I_{DCH}$, the rising-edge detector in Figure 8.5(b) sets for how long. When either $v_{O^+}$ or $v_{O^-}$ output turns high, it triggers, through $OR_{REF}$,
the rising-edge detector, which remains high for the designed 100-ns delay \( t_{DLY} \) that \( R_{R1}C_{R1} \), and \( R_{R2}C_{R2} \) set. Since the \( v_O^+ \) or \( v_O^- \) outputs are always low before \( CP_{REF} \) is enabled, the inverted and delayed input to \( AND_R \) is initially high, so that when either \( v_O^+ \) or \( v_O^- \) turns high, for the time that the RC network allows, the rising-edge detector’s output also stays high. Therefore, if \( v_O^+ \) transitions to a high state, for example, logic gate \( NAND_{CH} \) trips and engages \( MP_{CH} \) until the delayed signal, also fed into \( NAND_{CH} \), changes to a low state. Conversely, \( AND_{DCH} \) engages \( MN_{DCH} \) when \( v_O^- \) rises. The rise-edge detector’s series RC network maintains the signal high for \( t_{DLY} \), which means \( C_{REF} \) charges or discharges for approximately 100 ns. Note that a constant delay fixes \( C_{REF} \)’s charge variation \( \Delta Q_{REF} \) to

\[
\Delta Q_{REF} = I_{CH}t_{DLY} = I_{DCH}t_{DLY},
\]  

independent of \( C_{REF} \), which only influences voltage change \( \Delta v_{REF} \) (Equation 8.1). The precharge local bias block that only operates during precharge and biases \( CP_{VC} \) and \( CP_{SW} \) also generates \( I_{CH} \) and \( I_{DCH} \) and powers \( CP_{REF} \), so the tuning reference circuit only dissipates quiescent power during a diminutive fraction of every vibration cycle.

Leakage currents in the circuit and PCB board, however, discharge \( C_{REF} \) when the precharger is off, for about 33 ms with 30-Hz vibrations. This means \( v_{REF} \) droops
between sampling events and to ensure $v_{\text{REF}}$ can increase $\Delta Q_{\text{REF}}$ must, therefore, surpass leaked charge $Q_{\text{LEAK}}$:

$$\Delta Q_{\text{REF}} > Q_{\text{LEAK}} \Rightarrow I_{CH}t_{\text{DLY}} > I_{\text{LEAK}}T_{\text{VIB}}. \quad (8.3)$$

As a result, leakage current effectively decreases $\Delta v_{\text{REF}}$ when $v_{\text{REF}}$ tries to increase to $V_{\text{IBLEAK}}$

$$\Delta v_{\text{REF,UP}} = \frac{\Delta Q_{\text{REF}}}{C_{\text{REF}}} - I_{\text{LEAK}}T_{\text{VIB}}. \quad (8.4)$$

However, when $v_{\text{REF}}$ tries to decrease, $I_{\text{LEAK}}$ improves its change to $V_{\text{IBLEAK}}$

$$\Delta v_{\text{REF,DOWN}} = -\frac{\Delta Q_{\text{REF}}}{C_{\text{REF}}} - I_{\text{LEAK}}T_{\text{VIB}}. \quad (8.5)$$

For this reason, while at steady state, $C_{\text{REF}}$ raises $v_{\text{REF}}$ several steps for each time $C_{\text{REF}}$ decreases $v_{\text{REF}}$. Notice, by the way, $\Delta Q_{\text{REF}}$ and $Q_{\text{LEAK}}$ do not depend on $C_{\text{REF}}$ so increasing $C_{\text{REF}}$ does not help.

### 8.2.2 Latch Comparator

After each precharge event, enabling signal $v_{\text{LATCH}}$ closes $MN_{\text{EN}}$ and opens $MP_{\text{EN1-EN4}}$ in Figure 8.6 to engage $CP_{\text{LATCH}}$. The complementary outputs of buffer preamplifier $AMP_{\text{PRE}}$ create a current imbalance in differential transistors $MN_{\text{11}}$ and $MN_{\text{12}}$ that triggers the positive feedback loop across $MN_{\text{13-14}}$ and $MP_{\text{15-16}}$ and drives complementary output inverters $MN_{\text{2A-MP2A}}$ and $MN_{\text{2B-MP2B}}$. Once enabled, nodes $v_{\text{13}}$ and $v_{\text{14}}$ latch to supply or ground, ensuring the circuit remains in a zero-current state after reaching a decision to minimize energy consumption [206]. Preamplifier $AMP_{\text{PRE}}$ drives signals within $CP_{\text{LATCH}}$’s input common-mode range (ICMR), which requires its inputs to be sufficiently high to turn on transistors $MN_{\text{11}}$ and $MN_{\text{12}}$, and shunts switching noise that $CP_{\text{LATCH}}$ couples back into $v_{\text{INL+}}$ and $v_{\text{INL-}}$. What is more, $AMP_{\text{PRE}}$ increases
CP\textsubscript{LATCH}'s input overdrive to accelerate its response and avoid inadvertent transitions by amplifying the difference sensed in $v_C$ and $V\textsubscript{BAT}$ before feeding them into CP\textsubscript{LATCH}.

### 8.2.3 Preamplifier

To fully accommodate $v_C$'s range, from ground to above $V\textsubscript{BAT}$, and amplify enough of $V\textsubscript{BAT}$ and $v_C$'s difference for CP\textsubscript{LATCH} to operate properly, AMP\textsubscript{PRE} in Figure 8.7 features complementary n- and p-type differential pairs MP\textsubscript{2A}-MP\textsubscript{2B} and MN\textsubscript{4A}-MN\textsubscript{4B}. Source followers MN\textsubscript{1A}-MN\textsubscript{1B} level-shift the inputs to help input pair MN\textsubscript{4A}-MN\textsubscript{4B} maintain enough dynamic range across resistor load $R\textsubscript{L1}$-$R\textsubscript{L2}$ when $v_C$ exceeds $V\textsubscript{BAT}$. Architecturally, MN\textsubscript{4A}-MN\textsubscript{4B} feed currents directly to $R\textsubscript{L1}$-$R\textsubscript{L2}$ while MP\textsubscript{2A}-MP\textsubscript{2B} fold theirs into the load through cascodes MN\textsubscript{3A}-MN\textsubscript{3B}. As a result, outputs $v_P^+$ and $v_P^-$ swing between $V\textsubscript{BAT}$ and roughly 1 V below $V\textsubscript{BAT}$ with 16 µA into 62.5 kΩ, which is sufficiently high to drive CP\textsubscript{LATCH}'s input NMOS pair. Finally, AMP\textsubscript{PRE} derives its bias currents from the same local precharge bias-current generator as the charge pump, which.

Figure 8.6. Comparator CP\textsubscript{REF}'s latch CP\textsubscript{LATCH} (transistor dimensions are in µm).
the system only enables for a small fraction of each vibration period to keep quiescent losses low.

### 8.2.4 Modified Harvest Detection

As the system initially powers up, \( v_{\text{REF}} \) is initially low near ground and \( C_{\text{VAR}} \) undercharges. Consequently, when the harvest-detect block is enabled with signal \( v_{\text{HARV-START}} \), it immediately senses \( v_C \) below \( V_{\text{BAT}} \) and incorrectly triggers prematurely while \( C_{\text{VAR}} \) is still near \( C_{\text{MAX}} \), engaging additional subsequent reset and precharge phases that increase losses. To avoid any premature triggering of harvest-detect comparator \( CP_{\text{HARV-DET}} \) when \( C_{\text{VAR}} \) severely undercharges, the subsystem was modified as shown in Figure 8.8 to detect only when \( v_C \) falls below \( V_{\text{BAT}} \) from higher, not just when it is lower. To that end, \( CP_{\text{HARV-DET}} \) always starts high, which is the forced state when disabled, so that if

![Figure 8.7. Comparator CP\textsubscript{REF}’s preamplifier \textsubscript{AMP\textsubscript{PRE}} (transistor dimensions are in \textmu m).](image-url)
Figure 8.8. Modified harvest-detection subsystem.

$C_{\text{VAR}}$ is initially undercharged ($v_C < V_{\text{BAT}}$) the comparator remains high and there is no change at its output. As $v_C$ increases because $V_{\text{BAT}}$ charges $C_{\text{VAR}}$ through $MP_H$, $CP_{\text{HARV-DET}}$ trips low when $v_C$ reaches $V_{\text{BAT}}$, but the rise-detect logic remains low and does not set latch $SR_{\text{HARV-DET}}$. Note that under normal operating conditions, $v_C$ will precharge to $V_{\text{BAT}}$ and increase above as $i_{\text{HARV}}$ charges the battery, also causing the comparator output to drop. As $C_{\text{VAR}}$ approaches $C_{\text{MIN}}$, $i_{\text{HARV}}$ drops to 0 A, $v_C$ drops below again $V_{\text{BAT}}$, but now $CP_{\text{HARV-DET}}$ trips from a low state to a high one, causing the rise-edge detector to set the latch and open $MP_H$ by forcing $v_{\text{HARV}}$ high. As a result, $MP_H$ opens only at the end of the harvest phase, after $C_{\text{VAR}}$ truly reaches $C_{\text{MIN}}$.

Figure 8.9. Modified harvest-detect comparator $CP_{\text{HARV-DET}}$ with class AB output stage (transistor dimensions are in $\mu$m).
Comparator CP\textsubscript{HARV-DET} in Figure 8.9, however, trips low and high in each cycle and, as a result, both decisions must be sufficiently fast to avoid substantial shoot-through (or short-circuit) losses in the subsequent logic gates. For instance, Class A amplifier MP\textsubscript{5} is able of pulling node $v\text{O}_5$ high quickly, but will fall to a low state slowly, as it is slew-limited by bias current $I_{B5}$. A slow ramping signal causes any subsequent inverters to sink considerable currents from $V_{BAT}$ because both NMOS and PMOS transistors stay on simultaneously for an extended time. For this reason, class AB output stage in Figure 8.10 drives the output signal of CP\textsubscript{HARV-DET} [207]. If $v\text{O}_5$ is initially low and trips high (i.e., the “fast” transition), nodes $v\text{O}_6\text{A}$ and $v\text{O}_6\text{B}$ are initially high and $I_{B6A}$ biases $M\text{N}_6\text{A}$. As a result, a low-to-high input transition will quickly turn on $M\text{P}_7$ and off $M\text{N}_7$ simultaneously. Since the low-to-high $v\text{O}_5$ transition is not slew-limited, shoot-through currents through the $M\text{N}_7$-$M\text{P}_7$ inverter and subsequent stages are not significant.

However, now that $v\text{O}_5$ is high, $v_{DT}$ is also high and $M\text{P}_{DT2}$ shuts off $I_{B6A}$. Therefore, a subsequent high-to-low input transition will first turn off $M\text{P}_7$, and $M\text{N}_7$ will only turn on to discharge $v\text{O}_6$ to ground after the delay $C_{DT}$ creates to activate $I_{B6A}$ and

Figure 8.10. CP\textsubscript{HARV-DET}’s class AB output stage with dead time (transistor dimensions are in $\mu$m).
bias MN6A. Essentially, during the slew-limited $v_{os}$ transition there is a dead time, set by how fast $I_{BDT}$ discharges $C_{DT}$. Note also that switches MNHA and MNHB create hysteresis in $CP_{HARV-DET}$, where MNHA connects to output $v_{HV-END}$ to ensure there is hysteresis available when the comparator is first enabled and $v_{HV-END}$ is initially high. Also, the deglitch circuit in Figure 6.13 delays enable signals to avoid glitches when powering up.

### 8.2.5 Other Modifications

The precharger also features dead time that forces both MEP and MN D to remain off after energizing L and $C_{VAR}$. During this time, $v_{SW}$ falls to ground as fast as inductor current $i_L$ can discharge parasitic capacitance $C_{SW}$ at the switching node. In the event that the de-energizing step starts before $C_{SW}$ is fully discharged, $CP_{SW}$ can trip prematurely because $v_{SW}$ would be greater than ground. The result is MN D shuts off early and L de-energizes through its lossy body diode. For this reason, a 10-ns delay circuit was added to the precharge logic that delays $CP_{SW}$’s enabling signal. This is enough time since, even at the worst case with the highest $V_{BAT}$ (4.2 V), a conservative low current estimate of 20 mA can discharge up to 47.6 pF in 10 ns. Capacitor $C_{SW}$ is expected to be considerably lower.

Other modifications to the system improved the robustness of the IC. For instance, more capacitance was added to precharge-detect comparator $CP_{PCH-DET}$’s deglitch circuit and its differential pair current was increased to 1 nA since the common-mode source node of the differential pair was slow to charge at power up, causing the comparator to trip early under some conditions. In addition, a 100-ns digital delay block separates the end of precharge from when $CP_{REF}$ enables to compare $v_C$ and $V_{BAT}$. In this way, $C_{VAR}$ settles to its final voltage before measuring it.
8.2.6 Test Mode, ESD Protection, and Layout

The IC also features a test mode that allows each system component to be enabled and evaluated individually, including comparator CPREF and charge pump currents IC and IC. The system also includes an operational mode that bypasses and disables the

Figure 8.11. Floorplan of the 1.5 x 1.5 mm² self-tuning harvester IC.

Figure 8.12. Layout diagram of the 1.5 x 1.5 mm² self-tuning harvester IC.
self-tuning precharger, so that the IC functions with a fixed $V_{REF}$. Internal pin-out buffers drive digital signals important to the functionality of the system, such as $V_{HARV}$, $V_{PCH}$, $V_{GN}$, $V_{GP}$, $V_{HV-END}$, $V_{E-END}$, $V_{D-END}$, and $CP_{REF}$’s outputs $V_{O^+}$ and $V_{O^-}$, off chip to allow debugging and experimental evaluation.

The circuit in Figure 6.19(a) protects all pins, except the $V_{SW}$ and $V_{REF}$ nodes, from electrostatic discharge (ESD) events. By avoiding connecting any ESD circuitry to $V_{SW}$ and $V_{REF}$ nodes guards them from added parasitics and leakage currents that can potentially affect the system’s behavior. Self-adjusting reference voltage $V_{REF}$, for instance, was pinned out only for testing purposes, but no ESD protection was included to keep the large ESD circuit from leaking $C_{REF}$. Node $V_{REF}$, however, features a small local gate clamp like the one in Figure 6.19(b) to protect $C_{REF}$’s polysilicon plates. Off-chip OPA2354 analog amplifier buffers $V_{REF}$ to permit measurement.

In conclusion, the presented self-tuning energy harvester IC was implemented in a 0.7-µm BiCMOS process, as seen in the floorplan in Figure 8.11 and the layout diagram

Figure 8.13. Die photograph of the 1.5 x 1.5 mm$^2$ self-tuning harvester IC.
in Figure 8.12. The IC features 32 pins, but most are only used for measurements and experimental testing (e.g., buffered digital signals and test-mode setting inputs). The entire system includes test-mode logic, pin-out digital buffers, testpads, ESD protection circuits, $R_{\text{DLY}}$, and all bondpads ($80 \times 112 \ \mu m^2$) and was laid out inside the $1.5 \times 1.5 \ mm^2$ silicon area shown in Figure 8.12.

### 8.3 Experimental Prototype Results

The 0.7-µm BiCMOS $1.5 \times 1.5 \ mm^2$ silicon die pictured in Figure 8.13 integrates the proposed self-tuning energy-harvesting system. A 32-pin 7 x 7 mm² plastic quad-flat package (PQFP) encapsulates the IC, which includes the entire system, as presented in Figure 8.3, except $L$, $C_{\text{VAR}}$, and for testing purposes, bias, and sense resistors $R_{\text{PTAT}}$, $R_{\text{CTAT}}$, and $R_{\text{HARV}}$. Precharge-detect delay resistor $R_{\text{DLY}}$ (42 MΩ) is integrated completely on the silicon die. The IC also includes test-mode logic and pin-out digital buffers and was tested with the PCB in Figure 8.14. Off-chip OPA2354 buffers (same as OPA2357 but in different package) sense $V_{\text{BAT}}$, $V_{\text{REF}}$, $V_{\text{SW}}$, and $V_C$ as in Figure 8.15, while the $i_{\text{HARV}}$-
sensing circuit from Figure 7.6(a) was moved to the \( V_{BAT} \) side of the harvesting switch. Note that \( R_{HARV} \) was increased to 200 kΩ, but the instrumentation amplifier still senses the current through 100 kΩ of \( R_{HARV} \). A \( 1 \times 2 \times 2\text{-mm}^3 \) 10-µH Coilcraft inductor with a maximum equivalent series resistance (ESR) of 1 Ω served as the precharge inductor. The prototyped variable capacitor discussed in Chapter 7 served as the harvesting transducer, which resonates at 30 Hz between 991.2 and 156.8 pF when re-evaluated for the self-tuning harvester IC measurements.

As the experimental results from Figure 8.16 show, \( C_{VAR} \) generates, on average, up to 505.3 nA (\( i_{HARV} \)) when clamped to a 3.5-V battery. Switch \( MP_H \) conducts \( i_{HARV} \) into the battery, which when integrated over time, represents an average gain of 10.1 nJ/cycle. At the end of each harvesting phase, \( MP_H \) disengages and \( i_{HARV} \) drops to zero, and the reset phase follows with \( v_C \) gradually dropping to a minimum. The harvesting detection
circuit, which is active through the harvesting phase for roughly 17.77 ms/cycle (on average), consumes a measured quiescent current $I_Q$ of 2.63–3.75 nA, resulting in 209.76 pJ/cycle of used energy. Similarly, the precharge detector draws a measured $I_Q$ of 1.80–3.69 nA for the duration of the reset phase, for approximately 15.56 ms/cycle (on average), resulting in roughly 141.06 pJ/cycle. The nanoampere current generator, which biases both detection blocks, remains operational through the entire period for 33.33 ms, which corresponds to 30-Hz vibrations, sinks 2.48–2.96 nA from the 3.5-V supply, and uses an average of 320.34 pJ/cycle.

The system invests the necessary energy from the battery through L to charge $C_{VAR}$ to its 3.5-V target during every precharge phase. The self-tuning precharger energized L and $C_{VAR}$ (on average) for about 134.2 ns, producing a peak inductor current of 24.15 mA. Inductor L then de-energized into $C_{VAR}$ in 92.55 ns, resulting in an average invested energy of 6.72 nJ/cycle. That is an investment efficiency at 3.5 V of 90.3 % since the ideal investment (to 991.2 pF maximum capacitance) is 6.07 nJ. The precharge

Figure 8.16. Experimental measurements showing variable capacitor voltage $v_C$, harvesting current $i_{HARV}$, and extrapolated energy gain $E_{HARV}$. 
control circuit, which includes the zero-current sensor CP_{SW} and CP_{VC} that sets energizing time t_E, only operates during the energize and de-energize steps and use 44.19 pJ/cycle. The precharge bias generator powers when C_{VAR} reaches C_{MAX} to become functional after roughly 245.25 ns, after which the energize/de-energize sequence initiates. As a result, the generator uses 31.82 pJ/cycle, totaling the energy lost in the control circuit to 76.01 pJ/cycle.

Figure 8.17. Experimental waveforms showing variable capacitor voltage v_C and variable reference voltage v_{REF}.

Figure 8.18. Variable reference voltage v_{REF} during startup and through steady state.
Reference voltage $v_{\text{REF}}$, which sets $v_{\text{C}}$’s energizing time $t_{\text{E}}$, adjusts after each precharge phase and varies between 2 and 2.5 V when tested at 3.5 V, as seen in Figure 8.17. On average, the system raises $v_{\text{REF}}$ by 189.50 mV and decreases it by 164.38 mV by charging or discharging $C_{\text{REF}}$ (100 pF). An average of 376.33 pA leaks $C_{\text{REF}}$ to decrease $v_{\text{REF}}$ by 125.43 mV every cycle, limiting the rise in $v_{\text{REF}}$ to 64.07 mV and increasing the drops to 289.81 mV, as summarized in Table 8.1. For this reason, $v_{\text{REF}}$ increases on average 3.48 times for every time it decreases. Note, however, the off-chip test buffer used to measure $v_{\text{REF}}$ leaked considerable charge from $C_{\text{REF}}$. On average, though, each charge event in $C_{\text{REF}}$ uses 48.41 pJ/cycle, and the charge pump and $CP_{\text{REF}}$ power with the precharge comparators to dissipate 33.26–35.43 µA and 39.79–44.98 µA for 489.95 ns and use 57.11 pJ/cycle and 70.29 pJ/cycle, respectively. When the system first powers up, as shown in Figure 8.18, $v_{\text{REF}}$ charges incrementally each cycle from ground until it reaches steady state after about 25.63 cycles (on average).

Table 8.1. Summary of the self-tuning harvester IC performance.

<table>
<thead>
<tr>
<th>Die Information</th>
<th>1.5 × 1.5 mm² 0.7-µm BiCMOS IC</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of Transistors</td>
<td>977 Transistors</td>
</tr>
<tr>
<td>$V_{\text{BAT}}$ Range</td>
<td>2.7 V</td>
</tr>
<tr>
<td>Precharger</td>
<td>Efficiency</td>
</tr>
<tr>
<td>$CP_{\text{REF}}$</td>
<td>$I_{\text{O}}$</td>
</tr>
<tr>
<td></td>
<td>$v_{\text{O}}^+$ Delay</td>
</tr>
<tr>
<td></td>
<td>$v_{\text{O}}^-$ Delay</td>
</tr>
<tr>
<td>Charge Pump</td>
<td>$I_{\text{O}}$</td>
</tr>
<tr>
<td>Time ON</td>
<td>$t_{\text{ON,AVG}}$</td>
</tr>
<tr>
<td>Variable $v_{\text{REF}}$ (Averages)</td>
<td>$V_{\text{REF,MAX}}$</td>
</tr>
<tr>
<td></td>
<td>$V_{\text{REF,MIN}}$</td>
</tr>
<tr>
<td></td>
<td>$\Delta V_{\text{REF,UP}}$</td>
</tr>
<tr>
<td></td>
<td>$\Delta V_{\text{REF,DOWN}}$</td>
</tr>
<tr>
<td></td>
<td>$\Delta V_{\text{REF,LEAK}}$</td>
</tr>
</tbody>
</table>
Ultimately, the total energy the system drew from vibrations in $C_{VAR}$ exceeded all losses, producing a net gain of 2.434 nJ/cycle for a 3.5-V battery, which equivalent to 73.02 nW at 30 Hz. The system also produced gains of 1.930 and 3.885 nJ/cycle at 2.7 and 4.2 V, for 57.89 and 116.55 nW at 30 Hz, respectively, as summarized in Table 8.2.

Note that charging an actual battery is impractical during testing because the large capacities that commercial batteries feature lead to months long charge times. Instead, the 1-µF ceramic capacitor $C_{BAT}$ the harvester charged from 2.7 to 4.2 V in Figure 8.19 emulates a microscale battery.

Across 8 samples and 51 measurements, the harvester charged $C_{BAT}$ from 2.7 to 4.2 V (that is, gain 5.175 µJ) in 68.84 s, on average. This represents an average of 75.18 nW for the entire voltage range. Tuning reference $v_{REF}$ increased from 1.5 to 2.7 V, adapting itself to $V_{BAT}$. A fixed 2.3-V reference, which is the average value of $v_{REF}$ when $V_{BAT}$ crosses 3.5 V, results in less gain and extends $C_{BAT}$’s charge time, as seen in Figure

<table>
<thead>
<tr>
<th>Phase</th>
<th>Measured Energy (nJ/cycle)</th>
<th>$V_{BAT} = 2.7$ V</th>
<th>$V_{BAT} = 3.5$ V</th>
<th>$V_{BAT} = 4.2$ V</th>
</tr>
</thead>
<tbody>
<tr>
<td>Harvest Phase</td>
<td>Harvested Energy</td>
<td>+6.842</td>
<td>+10.073</td>
<td>+14.335</td>
</tr>
<tr>
<td></td>
<td>Control Circuits</td>
<td>-0.174</td>
<td>-0.210</td>
<td>-0.252</td>
</tr>
<tr>
<td>Reset Phase</td>
<td>Control Circuits</td>
<td>-0.102</td>
<td>-0.141</td>
<td>-0.171</td>
</tr>
<tr>
<td></td>
<td>Control Circuits</td>
<td>-0.058</td>
<td>-0.076</td>
<td>-0.090</td>
</tr>
<tr>
<td>Tuning Reference</td>
<td>$C_{REF}$ Avg. Charge</td>
<td>-0.032</td>
<td>-0.048</td>
<td>-0.065</td>
</tr>
<tr>
<td></td>
<td>Control Circuits</td>
<td>-0.095</td>
<td>-0.127</td>
<td>-0.156</td>
</tr>
<tr>
<td>nA Bias-CURRENT Generator</td>
<td></td>
<td>-0.245</td>
<td>-0.320</td>
<td>-0.391</td>
</tr>
<tr>
<td><em>Net Energy Gain (per cycle)</em></td>
<td></td>
<td>+1.930 nJ</td>
<td>+2.434 nJ</td>
<td>+3.885 nJ</td>
</tr>
<tr>
<td><em>Power Gain (at 30 Hz)</em></td>
<td></td>
<td>+57.89 nW</td>
<td>+73.02 nW</td>
<td>+116.55 nW</td>
</tr>
</tbody>
</table>

Table 8.2. Measured energy consumed and gained by the self-tuning harvester IC.
In other words, the harvester only generates 3.499 µJ to charge $C_{BAT}$ to 3.78 V in the same time frame the proposed circuit charged $C_{BAT}$ to 4.2 V. Note that the variable reference block is disabled during this latter experiment to avoid dissipating the power a fixed reference would not. In the end, the self-tuning $v_{REF}$ loop leads to a 47.9% improvement. Note this improvement is achieved even when ignoring the additional power a fixed internal reference would demand and, therefore, the actual improvement is potentially higher.

**8.4 Discussion**

With the self-tuning prototype IC, vibrations produced gains of 1.930, 2.434, and 3.885 nJ/cycle when $V_{BAT}$ is at 2.7, 3.5, and 4.2 V, respectively, generating 57.89, 73.02, and 116.55 nW with 30 Hz vibrations. By duty-cycling a microsensor system’s high-power functions to operate only when enough energy is harvested and stored in the battery, allow the system to extend its life. For instance, making the same load assumptions from Chapter 7, which permit 10 ms for 5 mW wireless telemetry (5 ms for
transmission and 5 ms for reception) and sensing at 10 µW for 1 ms, results in the system spending about 50 µJ. Then, the self-tuning harvester would replenish the system’s consumption in 863.88 s (14.4 min.), 684.88 s (11.4 min.), and 429.09 s (7.2 min.), as Table 8.3 summarizes. As a result, because of the energy gain vibrations produce, the wireless microsensor can sense and transmit only when the harvester recuperates the energy the system spends and, in this way, extend its operational lifetime.

As already mentioned, \( v_{REF} \) is prone to leakages. A digital counter and a conventional digital-to-analog converter (DAC) would avoid those effects, but at the expense of more silicon area. Based on the decision of \( CP_{REF} \), the digital counter would count either up or down a digital word that is converted to an analog voltage during the next precharge. Since the precharge reference voltage information is stored digitally until the next cycle, it is not affected by leakage currents. Also, \( v_{REF} \)’s accuracy and the energy investment it tunes would also improve if its incremental variation \( \Delta v_{REF} \) were proportional to difference \( v_C - V_{BAT} \) difference instead of being fixed, as is the case in the

<table>
<thead>
<tr>
<th>System Tasks</th>
<th>Duration</th>
<th>Power</th>
<th>Energy</th>
</tr>
</thead>
<tbody>
<tr>
<td>Transmission</td>
<td>5.00 ms</td>
<td>-5.00 mW</td>
<td>-25.00 µJ</td>
</tr>
<tr>
<td>Reception</td>
<td>5.00 ms</td>
<td>-5.00 mW</td>
<td>-25.00 µJ</td>
</tr>
<tr>
<td>Sensing</td>
<td>1.00 ms</td>
<td>-10.00 µW</td>
<td>-0.01 µJ</td>
</tr>
<tr>
<td>Total Consumption</td>
<td></td>
<td></td>
<td>-50.01 µJ</td>
</tr>
</tbody>
</table>

Harvester Requirement at 2.7 V: \( > 863.88 \) s +57.89 nW > +50.01 µJ
Harvester Requirement at 3.5 V: \( > 684.88 \) s +73.02 nW > +50.01 µJ
Harvester Requirement at 4.2 V: \( > 429.09 \) s +116.55 nW > +50.01 µJ
system presented. Notwithstanding, the prototyped implementation validates and
demonstrates the value of self-tuning the system to adapt to and compensate for a
constantly changing (i.e., charging and discharging) battery voltage, irrespective of
circuit non-idealities like losses, delays, offsets, etc.

8.5 Summary

The presented IC gains 1.930, 2.434, and 3.885 nJ/cycle from 30-Hz vibrations to
generate 57.89, 73.02, and 116.55 nW at battery voltages 2.7, 3.5, and 4.2 V and charges
a 1-µF battery-emulating capacitor from 2.7 to 4.2 V in 68.84 s. The system did this by
automatically tuning the energizing time of an energy-transfer inductor in finite and
constant steps, as determined by $\Delta V_{REF}$, to optimally precondition and precharge $C_{VAR}$ to
$V_{BAT}$ every cycle, irrespective of $V_{BAT}$. In this way, the system adjusts the energy
invested to what is needed, no more and no less. As a result, the energy harvester
produces a net energy gain that can extend the life of a self-powered microsystem
indefinitely.
CHAPTER 9

FINAL SUMMARY AND CONCLUSIONS

As electronic systems continue to reduce in size, microelectronic technology has reached the point where ultra-compact and autonomous devices like wireless microsensors are possible. These microscale devices provide sensing and monitoring functions from typically inaccessible environments, unobtrusively and in numerous quantities. Their sheer diminutive dimensions, however, render the upkeep of the on-board energy supplies impractical and demand that the device sustains itself and survive from the energy it stored initially. The result is that, because of the low energy densities typical microscale energy sources feature, the self-powered device only sustains operations for a short life. The environment, however, is a rich source of energy that can be harnessed to replenish continuously the microsystem’s energy supply and overcome the short-life shortcomings that plague such self-powered applications.

The primary objective of this research was to investigate and develop an electrostatic energy-harvesting voltage-constrained CMOS/BiCMOS integrated circuit (IC) that harnesses ambient kinetic energy from vibrations with a vibration-sensitive variable capacitor and channels the extracted energy to charge an energy-storage device (e.g., battery). To that end, the research developed an energy-harvesting system that synchronizes to variable capacitor $C_{\text{VAR}}$’s state as it cycles between $C_{\text{MAX}}$ and $C_{\text{MIN}}$ by controlling each functional phase of the harvester and adjusting to different voltages of the on-board battery. One of the major challenges of the system was performing all of these duties without dissipating the energy harnessed and gained from the environment,
which was achieved by employing low-energy dissipation strategies and time-managing circuit operation. The energy-harvesting system prototypes were analyzed, designed, developed, and validated using Texas Instruments’ high-volume 0.7-µm BiCMOS process and Table 9.1 summarizes the performance of each. This chapter reviews and summarizes the conclusions reached and contributions derived from the presented work, including important tradeoffs and future potential developments.

### Table 9.1. Summary of system-level performance of the tested prototype ICs.

<table>
<thead>
<tr>
<th>Prototype</th>
<th>Parameter</th>
<th>$V_{BAT}$ Range</th>
<th>2.7 V</th>
<th>3.5 V</th>
<th>4.2 V</th>
</tr>
</thead>
<tbody>
<tr>
<td>Fixed-Reference Harvester IC Prototype</td>
<td>Net Energy Gain (per cycle)</td>
<td></td>
<td>+1.273 nJ</td>
<td>+2.140 nJ</td>
<td>+2.869 nJ</td>
</tr>
<tr>
<td></td>
<td>Power Gain (at 30 Hz)</td>
<td></td>
<td>+38.19 nW</td>
<td>+64.20 nW</td>
<td>+86.07 nW</td>
</tr>
<tr>
<td></td>
<td>50-µJ Recharge Time</td>
<td></td>
<td>&gt; 1309.51 s</td>
<td>&gt; 778.97 s</td>
<td>&gt; 581.04 s</td>
</tr>
<tr>
<td>Self-Tuning Harvester IC Prototype</td>
<td>Net Energy Gain (per cycle)</td>
<td></td>
<td>+1.930 nJ</td>
<td>+2.434 nJ</td>
<td>+3.885 nJ</td>
</tr>
<tr>
<td></td>
<td>Power Gain (at 30 Hz)</td>
<td></td>
<td>+57.89 nW</td>
<td>+73.02 nW</td>
<td>+116.55 nW</td>
</tr>
<tr>
<td></td>
<td>50-µJ Recharge Time</td>
<td></td>
<td>&gt; 863.88 s</td>
<td>&gt; 684.88 s</td>
<td>&gt; 429.09 s</td>
</tr>
</tbody>
</table>

9.1 Conclusions

Harnessing ambient energy to charge and replenish a wireless microsystem’s energy supply, such as a thin-film Li Ion, breaks the dependence between its operational lifetime and the initial energy stored on reservoir. Energy in the nearby environment around the system is available is different forms, mainly as light, heat-generated temperature gradients, electromagnetic radiation, and mechanical movements and vibrations. Extracting each type of ambient energy source presents different microscale
integration challenges, and the power levels that can be achieved strongly depend on the specific conditions surrounding the application. This research focused on harvesting kinetic energy from vibrations because these are present and plentiful in a wide variety of environments and applications.

Kinetic energy from vibrations can be harnessed by using magnetic fields (electromagnetic), electric fields (electrostatic), or strain on a piezoelectric material. One key disadvantage of piezoelectric or electromagnetic harvesters is that they require the integration and miniaturization of piezoelectric or magnetic materials, which are foreign to conventional silicon-based processes. Electrostatic harvesters, on the other hand, require a microscale variable capacitor, which is compatible with present-day mainstream and mature MEMS technologies, as evidenced in current accelerometer products. Even then, they require an initial charge to exist in the capacitor as an investment to initiate the harvesting process. This investment must be a fraction of the energy gained and must synchronize to the vibration source by precharging $C_{VAR}$ when its plates are closest together, or at $C_{MAX}$. By either holding the capacitor’s voltage or charge constant as vibrations force $C_{VAR}$’s plates apart (and capacitance decreases) harnesses kinetic energy and converts it to electrical. Nevertheless, unlike charge-constrained systems, holding voltage constant maintains system voltages low, below the limits of mainstream low-cost CMOS or BiCMOS semiconductor processes.

To hold voltage constant without adding another source or device, the presented harvester clamps $C_{VAR}$ with the already-existing microsystem’s battery, which is the intended target to store the harvested energy anyways. In this way, vibrations charge directly the battery with the generated harvesting current. First, however, the battery
makes the necessary and required investment to precharge the capacitor to battery voltage $V_{\text{BAT}}$ when its capacitance is highest ($C_{\text{MAX}}$), and therefore, establish the electrostatic force against which vibrations will work. Connecting the battery to the precharged capacitor afterwards generates, as $C_{\text{VAR}}$’s plates separate and capacitance decreases, the harvesting current that directly charges the microsystem’s energy supply. When the capacitor reaches its minimum value, to avoid discharging the battery, the system then disconnects $C_{\text{VAR}}$ and keeps it open-circuited, until vibrations force it to $C_{\text{MAX}}$ and the process repeats again. As long as the battery gains more energy than what it losses by investing to precharge $C_{\text{VAR}}$, the energy available in the self-powered microsystem increases and replenishes what it consumes. Additional circuits, however, must not only detect when to precharge $C_{\text{VAR}}$ and disconnect it from the battery at the end of harvesting, but also direct and control the precharge process. Yet, any supplementary synchronizing circuitry must dissipate as little energy as possible. In other words, the energy harvested must surmount not only the initial investment but also the losses associated with all the monitoring and control circuits in the system to guarantee a net energy gain.

Generating a net energy gain in the system to increase the battery’s state of charge demands and requires minimal dissipation from any necessary accompanying circuitry. For instance, the system transfers the required investment from the battery to the capacitor efficiently with an inductor. Although the inductor-based precharger is not for practical purposes truly “lossless” and does dissipate additional energy from the battery (which manifests as extra investment), it is a more efficient procedure than simply connecting the uncharged capacitor to the battery. Similarly, once precharged, connecting the capacitor to the battery through a transistor switch saves the added losses an
asynchronous diode would incur. Because precharge only lasts a diminutive fraction of each vibration cycle (for instance, about 250 ns out of each 33.3 ms cycle - that is less than 0.001 %), the presented energy-harvesting ICs save a significant amount of energy by turning off all precharge control circuits when unused, even though they sink relatively high bias currents in the microampere range.

However, monitoring circuits that detect when the capacitor reaches its minimum value (to detect the end of harvest) and maximum value (to detect when to precharge) must remain available and enabled during at least half of each cycle. Fortunately, since vibrations feature low frequencies, each detection block does not need to be fast and could operate with subthreshold currents in the nanoampere range. What is more, a time-discrete feedback loop, which only engages for a brief fraction of each cycle, adjusts the battery’s energy investment to ensure it does not over- or undercharge $C_{VAR}$ during precharge, and in this way, guarantee only the necessary investment is made every cycle.

In conclusion, the battery-constrained energy-harvesting system presented in this research achieves a net energy gain that can potentially replenish the needs of self-powered microsystems such as wireless microsensors. The system senses and synchronizes efficiently to the vibration-driven variable capacitor as it cycles from $C_{MAX}$ and $C_{MIN}$ by managing and biasing its circuits to operate only when needed and with just enough energy. Also, the system tunes the invested energy to what is necessary, no more and no less, regardless of the conditions of the system. Although energy is extracted at a slow rate, meaning that only low power levels can be achieved, duty-cycling the microsensor functions allows it to recharge the battery and operate only when enough energy has been accumulated. As a result, the energy harvester produces a net energy
gain that can extend the life of self-powered wireless microsensors indefinitely or at least limited only by the natural wear and tear of the device rather than its initial stored on-board energy.

9.2 Contributions

The primary contribution of this work was the design and development of two energy-harvesting integrated circuits that extract kinetic energy from ambient vibrations to charge an energy-storage device. At the heart of the proposed system lies a novel energy-harvesting scheme based on constraining the voltage of a mechanical vibration-driven variable capacitor with the device that will store the harnessed energy, or in other words, the system’s already-existing battery. In this way, the resulting harvesting current directly charges the target storage device.

To achieve a net energy gain, the system precharges, monitors, and synchronizes to the vibration-driven capacitor without dissipating all the energy extracted from the environment during each cycle. As a result, additional contributions include the development of low-energy circuit design strategies that minimize the impact the system has on energy gain. For instance, the research work developed an efficient inductor-based energy-transfer circuit to precharge the variable capacitor and a self-adapting control strategy that precharges it with the correct amount of energy regardless of the battery state, while only engaging when necessary for an extremely short fraction of each cycle. Similarly, the phase-monitoring blocks, which detect $C_{\text{MAX}}$ and $C_{\text{MIN}}$ during the reset and harvesting phases, turn on and engage only during each respective half-cycle, but since these operate for longer portions of time, they draw just enough energy to function by being biased in deep subthreshold. In this manner, all losses in the harvester are
minimized to below what is harvested every cycle, such that a net energy gain exists. The result of these contributions is an efficient energy-harvesting and battery-charging integrated circuit, which when combined with a vibration-driven transducer, in the form of a mechanical capacitor, can convert kinetic energy from vibrations to restock continuously a self-powered microsystem’s energy supply.

The presented research has generated several publications in peer-reviewed journals, conferences, industry periodicals, and trade web-magazines. Three papers have already been published in peer-reviewed journals, while is one is currently awaiting review. The research work has also been presented and published in professional conferences with proceedings and industry-based periodicals. Finally, several web-based articles have been published in industry web-magazines, including Power Management Design Line’s second most read article of 2006. The following list presents a comprehensive record of all the publications related to this research:

Peer-Reviewed Journals


Peer-Reviewed Conferences with Proceedings


Industry Periodicals


**Industry Web-Magazines**


### 9.3 Limitations

The major advantage that electrostatic harvesters have over other vibration harvesters, such as their piezoelectric and electromagnetic counterparts, is their compatibility with state-of-the-art MEMS-based capacitors, since no additional foreign
materials are required. However, they also suffer from several drawbacks, such as needing an energy investment. What is more, although constraining voltage with the battery permits compatibility with low-cost, low-voltage semiconductor processes, it restricts the amount of energy that could potentially be harvested. In essence, charge-constrained or higher-voltage-constrained harvesters can extract more energy per cycle by imposing greater electrostatic force on the capacitor. Additional limitations that are specific to the presented research work are further discussed in this section, including, among others, lack of adaptability to varying vibration frequencies and implementation of a microscale variable capacitor and thin-film battery.

9.3.1 Energy Investment and Synchronization Requirements

The main drawbacks of electrostatic harvesters, in general, are the requirements to invest energy every cycle and synchronize with the vibration-driven variable capacitor. The fact that an investment is required to precharge $C_{VAR}$ demands that the battery must have some initial energy to invest. In other words, the system cannot harvest energy if the battery is completely depleted, but as long as some energy is available to precharge $C_{VAR}$, energy will be harvested and accumulated in the battery. The energy-harvesting system must also synchronize to $C_{VAR}$ so that when at $C_{MAX}$ it precharges the capacitor before connecting to the battery and disconnects the capacitor when it reaches $C_{MIN}$, which requires monitoring and sensing circuitry.

9.3.2 Vibration Frequency Variability

The presented harvesting circuit was tested with 30-Hz vibrations, since this was the resonant frequency of the macroscale capacitor used for experimental evaluation of each prototype IC. The tested frequency lies within the range of typical ambient
vibrations, which feature their highest acceleration peaks at frequencies lower than 500 Hz [94]-[95]. However, even though each energy harvesting IC prototype can synchronize to $C_{VAR}$’s maximum-to-minimum variations, its energy use is not optimized to frequency changes. Designing the energy-harvesting system for lower frequencies would require more energy from the detection blocks (such as the precharge and harvest detection subsystems) and nanoampere bias-current generator, since these engage for longer times. In other words, each vibration cycle, and therefore, each half-cycle is longer. Note that the precharger still is engages for the same amount of time, as long as $C_{MAX}$ is the same.

Without counting the losses from the detection and nanoampere current generator blocks, the system in Table 8.2 gains 3.105 nJ/cycle at 3.5 V - that is, harvested energy minus investment, precharger losses, and variable reference losses. The detection blocks and nanoampere bias generator dissipate 671 pJ/cycle and can remain enabled for about 4.63 times longer before cancelling these gains, assuming $\Delta C_{VAR}$ remains unaffected at different frequencies and $R_{DLY}$ and $R_{HARV}$ are adjusted accordingly. Therefore, the system theoretically maintains a net energy gain down to 6.48-Hz vibrations. Note that if the detection blocks are designed to engage for a fraction of each half-cycle, however, could improve the system’s efficiency even at lower frequencies, where now only the nanoampere bias generator, which is always enabled, poses a limit. Conversely, higher frequencies require faster detection blocks, and therefore, higher bias currents, so that even though shorter turn-on times reduce their energy consumption, energy gains are still limited by the possibly higher bias currents required. The ideal solution, however, is that
the battery-constrained energy harvester adapts and self-optimizes to different vibration frequencies.

### 9.3.3 Microscale Variable Capacitor

Each energy-harvesting IC prototype was tested and evaluated with a non-optimal macroscale variable capacitor. The fabrication of an energy-harvesting MEMS-based variable capacitor prototype, ideally, is similar to existing state-of-the-art accelerometer products, although practically it presents many challenges. Optimizing for voltage-constrained harvesters requires that the capacitor maximizes its capacitance difference $\Delta C_{VAR}$, rather than its maximum-to-minimum ratio $C_{MAX}/C_{MIN}$, which benefits charge-constrained systems. To create sufficiently large capacitances, the device must feature relatively large areas and therefore has to cope with low pull-in voltages and stiction. As a result, a MEMS device will probably feature a smaller $\Delta C_{VAR}$ and a lower $C_{MAX}$ values (e.g., $\Delta C_{VAR} \approx 500–100$ pF) than those achieved with the macroscale prototype. Furthermore, since its total mass is significantly less, its resonance frequency will probably be higher. However, even though a smaller $\Delta C_{VAR}$ reduces the energy harvested, a lower $C_{MAX}$ requires less investment energy. A shorter period also decreases the time and energy each detection block and the nanoampere bias-current generator operate in each cycle. Yet, as stated previously, faster frequencies require faster comparators and adjustment of the precharge detection RC delay. Nevertheless, adapting a MEMS variable capacitor to harvesting applications is an area of active research, typically including capacitances ranging between 400 and 50 pF and even up to 1570 to 62 pF [97], [179], [181]-[182], [188]-[189].
Ideally, the device should resonate with frequencies lower than 500 Hz and respond to accelerations between 1-12 m/s², as expected from typical ambient vibrations [94]-[95]. Maximum capacitance $C_{\text{MAX}}$ must be greater than 500 pF to achieve a net energy gain of at least 500 pJ/cycle, if $C_{\text{MIN}}$ and $C_{\text{PAR}}$ are 100 pF and 10 pF, respectively. This assumes an investment efficiency of 90% and at least 922 pJ dissipated by the control and detection circuitry (based on the results from the 30-Hz self-tuning energy harvester IC at 3.5 V in Table 8.2). For instance, to precharge 510 pF ($C_{\text{MAX}}$ plus $C_{\text{PAR}}$), the system would invest 3.471 nJ and harvest 4.90 nJ, resulting in a net gain of 507 pJ/cycle. Note, however, that control losses depend on the vibration frequency. Finally, the capacitor’s pull-in voltage should be greater than 4.2 V, the highest voltage a Li Ion can sustain, as summarized in Table 9.2.

### 9.3.4 CMOS Compatibility

The presented IC prototypes were designed and fabricated using the BiCMOS process available to this research. However, as previously discussed, the designs only used 5-V NMOS and PMOS devices, and in only two instances were the bipolar features

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Requirements</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum Capacitance, $C_{\text{MAX}}$</td>
<td>$&gt; 500$ pF</td>
</tr>
<tr>
<td>Minimum Capacitance, $C_{\text{MIN}}$</td>
<td>$&lt; 100$ pF</td>
</tr>
<tr>
<td>Parasitic Capacitance, $C_{\text{PAR}}$</td>
<td>$&lt; 10$ pF</td>
</tr>
<tr>
<td>Resonant Frequency</td>
<td>$&lt; 500$ Hz</td>
</tr>
<tr>
<td>Driving Acceleration</td>
<td>1-12 m/s²</td>
</tr>
<tr>
<td>Pull-In Voltage</td>
<td>$&gt; 4.2$ V</td>
</tr>
</tbody>
</table>

Table 9.2. Requirements of a MEMS variable capacitor.
of the process exploited (as in the p-tank for CP\textsubscript{SW}’s input pair in Figure 6.6 and the vertical NPN BJTs for the precharge bias-current generator in Figure 6.7). Converting to an all-CMOS design is feasible by allowing CP\textsubscript{SW}’s input NMOS pair sources to connect directly to the substrate or using an additional step-up input buffer to lower the comparator’s input common-mode range and using subthreshold MOSFETs in place of BJTs in the precharge bias-current generator. However, such conversion comes at the expense of lower accuracy and larger areas, as adding an additional input pair increases offset while using subthreshold transistors to generate microampere currents require large width-to-length ratios. Nevertheless, this is not expected to be a significant obstacle to the development of an energy-harvesting system in a low-voltage CMOS process.

9.3.5 Battery

Because of practical constraints, the system was not tested with a real battery. Otherwise, given the low power levels harvested, it would take an impractical length of time to see any change in battery voltage. As an example, a 100-µAh battery is equivalent to a charge of 360 mC, or approximately a 103 mF capacitor at 3.5 V. This means that to measure just a 100 mV increase, the battery would need to gain 36.6 mJ. By harvesting 2.434 nJ/cycle (Table 8.2), this would take over 15 million vibrations cycles! Approximately 140 hours are necessary with 30-Hz vibrations. Running the experimental setup for such a long time raised reliability concerns, such as heating of the prototype capacitor and the shaker. Therefore, the energy harvester charges a 1-µF capacitor instead. It is important to note, nevertheless, that an actual thin-film battery must be able to supply the precharge current, which can be up to 30 mA, while also leaking little current to avoid cancelling out the gains of the harvester. State-of-the-art thin-film Li
Ions, which compromise an area of active research, are currently capable of 0.1-5 mAh capacities within areas less than 25 cm$^2$ and 10-15 µm thick vapor-deposited layers [57].

It is important to note that a Li ion battery is sensitive to overcharge and discharge, where if the cell voltage increases above 4.2 V, or decreases below 2.7 V, the battery will significantly degrade, and in some cases, even vent and explode. In charging a ceramic capacitor, the system circumvents the need for battery protection, which means a practical implementation requires additional energy to protect a thin-film Li Ion, for example. This extra energy, however, need not be substantial when duty-cycling the circuit to engage only for a fraction of the vibration cycle, every several cycles. In addition, for maximum battery life and capacity, the Li Ion ideally should be carefully charged following a constant-current/constant-voltage methodology. Nevertheless, in an energy-harvesting system it is more sensible simply to charge the battery as efficiently as possible with whatever energy is present. In this case, $i_{\text{HARV}}$ always charges $V_{\text{BAT}}$ when it is available.

### 9.3.6 Other Limitations

The presented IC prototypes occupied die areas of 1 x 1 mm$^2$ and 1.5 x 1.5 mm$^2$. These, however, included redundant components, testing circuitry, and unessential bondpads, which a more size-conscious design does not need. Therefore, it is possible for an energy-harvesting IC to occupy a smaller silicon area, even when including large MΩ-range resistors. In addition, the QFP packages used were relatively large (7 x 7 mm$^2$) to accommodate the high number of pins (32), most of which were only necessary for test and measurement. The combination of smaller die areas with less pins allows the system
to be packaged in significantly smaller leadframes, if not integrated directly to the same silicon as a MEMS variable capacitor.

Finally, a key disadvantage of biasing circuits with subthreshold currents in the nanoampere range is that at high temperatures, leakage currents can overcome them, rendering the system unusable. Possibly, this problem can be contained by making bias currents proportional to absolute temperature (PTAT), ensuring that as temperatures increase, bias currents always stay above leakage currents. This, however, comes at the expense of higher quiescent losses. This was not the focus of this research, so the system was not designed for or tested at high temperatures and it is unclear if a net energy gain is possible. Note that a PTAT-only current generator would eliminate the need for the significantly high-resistance $R_{\text{CTAT}}$ (570 MΩ), shown in Figure 6.17.

9.4 Future Direction

Energy harvesting for microscale self-powered applications is a relatively young and immature technology, and its research is still in its infancy, promising many new future implementations and improvements. Although the presented research showed and proved the viability of voltage-constrained harvesters, future designs that improve efficiency will lead to higher energy gains (for the same $\Delta C_{\text{VAR}}$ variation) and possibly break the synchronization requirements of the presented system with an asynchronous approach. What is more, the eventual integration of the MEMS capacitor with the controlling circuits on the same silicon or package will be a key step forward in the implementation of electrostatic harvesters. Eventually, multiple harvesters will be integrated into the self-powered microscale application, replenishing the system’s consumption from multiple sources within the same package.
9.4.1 Asynchronous Design

A diode-based asynchronous design can overcome the main drawback of electrostatic harvesters: that it needs to synchronize to $C_{\text{VAR}}$, as seen in the possible implementation in Figure 9.1. In this circuit, as $C_{\text{VAR}}$ decreases, its voltage increases until it is clamped by diode D$_2$ at the sum of the battery voltages, forcing harvesting current to charge top battery $V_{\text{BAT2}}$ and bottom battery $V_{\text{BAT1}}$. After reaching $C_{\text{MIN}}$, as capacitance increases, $C_{\text{VAR}}$’s voltage $v_C$ drops but is held to $V_{\text{BAT1}}$ by diode D$_1$, where the bottom battery precharges the capacitor. In this way, no additional circuits, other than D$_1$ and D$_2$, are needed to detect $C_{\text{MAX}}$ or $C_{\text{MIN}}$, and therefore, there are no optimization requirements for different vibration frequencies (except for $C_{\text{VAR}}$’s resonant frequency). Note, however, that $V_{\text{BAT1}}$ is investing energy through a lossy diode, and as a result, it does not experience any net gain since it invests what it gains from harvest. In other words, since $V_{\text{BAT1}}$ precharges $C_{\text{VAR}}$ while constraining its voltage as capacitance increases, it discharges the equivalent harvesting current to cancel any gains. On the other hand, $V_{\text{BAT2}}$ does gain energy, since it does not have to invest anything because it is

![Figure 9.1. Possible implementation of an asynchronous electrostatic energy harvester.](image)
disconnected from $C_{VAR}$ while capacitance increases. Consequently, circuits that monitor each battery are required to swap them every several vibration cycles to ensure both can gain energy, ideally maintaining both voltages at similar levels to avoid situations were the energy invested is excessive or insufficient. What is more, by necessitating two batteries, the asynchronous harvester in Figure 9.1 complicates integration, although $V_{BAT2}$ could be replaced with a capacitor that discharges, although not completely, to the bottom battery every once in a while.

9.4.2 SiP Integration

An important future objective of energy harvesters is integration into the small self-powered microsystems they energize. One possibility will be to incorporate a fully co-packaged, microscale system-in-package (SiP) solution that confines the entire system within a small volume, which is a key feature of non-invasive low-cost sensing nodes in a wireless network. The SiP vision of the system compromises the energy-harvesting and battery-charging IC, a single in-package precharge inductor, an integrated thin-film Li Ion, and a MEMS vibration-sensitive mechanical variable capacitor, as Figure 9.2(a) illustrates. The thin-film Li Ion is deposited along the harvesting IC, while the inductor

![System-in-package (SiP) integration goal incorporating (a) all energy-harvesting system components within (b) a single package.](image)

Figure 9.2. System-in-package (SiP) integration goal incorporating (a) all energy-harvesting system components within (b) a single package.
and variable capacitor occupy different dies, each connected through bondwires and co-
packaged, as seen in Figure 9.2(b), in a planar or three-dimensional configuration.

The system could also be implemented within a microscale printed circuit board
(PCB), as shown in Figure 9.3, which possibly represents a feasible near-term solution
that uses current technologies. For instance, a state-of-the-art off-chip precharge inductor
can be as small as 1 x 2 x 2 mm$^3$ (Coilcraft’s EPL2010), while an off-chip polymer Li Ion
can be as thin as 500 µm (PowerStream PGE series). The energy-harvesting IC could be
packaged within a quad flat no-leads (QFN) package with footprint dimensions as small
as 3 x 3 mm$^2$ (for 8 leads). The MEMS device can be compatible with CMOS and
BiCMOS processes, and thus, it will be possible to integrate it to the same silicon die or
as a separate package.

**9.4.3 Multi-Harvester System**

Eventually, harvesting technology will mature sufficiently to integrate several
harvesters along with the wireless microsensor and thin-film Li Ion into a single self-
contained SiP solution. In this way, the thin-film lithium-ion microbattery stores the
energy from not only vibrations but also other sources of ambient energy sources, such as
light and thermal gradients. In such a system, as the example illustrated in Figure 9.4, an

![Figure 9.3. Integration of the energy-harvesting system at the board level.](image-url)
external photovoltaic panel wraps the system package, directly exposed to any light available. Part of the light energy is lost as heat, especially when outdoors, which raises the temperature of the panel and establishes a high temperature on top of the thermopiles underneath, while the cold junction of the thermoelectric layer is connected to the printed circuit board ground. Additionally, a MEMS variable capacitor harvests energy from vibrations through electrostatic means, with the presented energy-harvesting circuit. Each harvesting source collaborates and complements each other to provide sufficient energy for the system to operate. Other system components, such as the thin-film lithium-ion microbattery and supplementary circuitry, occupy separate substrates. The microelectronic circuit layer houses the control circuits of the harvesters, the microsensor interface electronics, and the battery charger that combines each energy source, plus any other safety function. Each substrate is interconnected with one another through the package’s leadframe and respective wirebonds, all inside a single plastic package, as conceptually illustrated in Figure 9.4.
9.5 Summary

This research developed, designed, built, and evaluated an energy-harvesting, battery-charging CMOS/BiCMOS system that scavenges ambient kinetic energy from vibrations with a mechanical variable capacitor to charge an energy-storage device. In particular, the presented system employs a novel energy-harvesting scheme based on constraining the voltage of the mechanical variable capacitor with the system’s already-existing battery. The system monitors and synchronizes to the vibration-driven capacitor without dissipating all the energy extracted during each vibration cycle and does this by only powering circuits when necessary and with just enough energy to function. The proposed harvester circuit surmounts the limits of finite energy sources in self-powered microdevices by extracting energy from the environment to create a net gain in the electrical system, which is continuously stored and accumulated in the system’s supply. The harnessed energy continuously replenishes the consumption of self-powered micro-applications, such as wireless microsensors, resulting in a theoretically infinite and inexhaustible supply, and, consequently, long-lasting operational lifetime not limited by the initial amount of energy stored.
APPENDIX A

DERIVATION OF INDUCTOR-BASED PRECHARGER VOLTAGE AND CURRENT EQUATIONS

Electrostatic energy harvesters require a charged motion-driven variable capacitor to harness ambient kinetic energy. In this way, as environment vibrations push capacitor $C_{VAR}$’s plates apart, capacitance decreases, generating harvesting current $i_{HARV}$ that charges the voltage-constraining battery $V_{BAT}$. As a result, the battery must invest the necessary energy to precharge $C_{VAR}$ to $V_{BAT}$ when at maximum capacitance $C_{MAX}$ (in other words, when $C_{VAR}$’s plates are closest together). To avoid overcoming and cancelling the harvesting gains with added electrical losses, an inductor-based circuit

![Diagram of four- and two-switch inductor-based prechargers.](image)

Figure A.1. (a) Four- and (b) two-switch inductor-based prechargers.
transfers the investment energy from $V_{BAT}$ to $C_{VAR}$ when at $C_{MAX}$, precharging the capacitor with high efficiency, as seen in the four- and two-switch implementations in Figure A.1.

**A.1 Four-Switch Precharger**

During each vibration cycle, $C_{VAR}$ is precharged to $V_{BAT}$ when it reaches $C_{MAX}$. This process is decomposed into an energize and de-energize sequence, where first $V_{BAT}$ energizes inductor $L$ and then de-energizes the accumulated energy in $L$ to $C_{VAR}$, as illustrated in Figure A.2 for the four-switch precharger circuit. The battery energizes $L$ by imposing $V_{BAT}$ across it when switches $S_{E1}$ and $S_{E2}$ are closed, as Figure A.2(a) illustrates. Since $V_{BAT}$ is relatively constant during each precharge, inductor current $i_L$ increases linearly,

$$i_L(t) = i_{BAT}(t) = \frac{V_{BAT}}{L} t, \quad (A.1)$$

until $L$ stores enough investment energy, at which point $S_{E1}$ and $S_{E2}$ open. After a dead time, where all switches are open, $S_{D1}$ and $S_{D2}$ close and deliver the stored energy to $C_{VAR}$.

---

**Figure A.2.** (a) Energize and (b) de-energize connections of the four-switch precharger.
When L de-energizes to \(C_{VAR}\), as shown in Figure A.2(b), the precharger is essentially an LC resonant circuit, which can be described by the following second-order differential equation

\[
\frac{d^2v_C}{dt^2} + \frac{v_C}{LC_{\text{MAX}}} = 0,
\]

(A.2)

where \(v_C\) is the capacitor’s voltage and the general solution is

\[
v_C(t) = K_1 \cos(\omega_{LC} t) + K_2 \sin(\omega_{LC} t),
\]

(A.3)

where \(K_1\) and \(K_2\) are constant values determined by the initial conditions of the circuit and \(\omega_{LC}\) is the circuit’s resonant frequency

\[
\omega_{LC} = \frac{1}{\sqrt{LC_{\text{MAX}}}}.
\]

(A.4)

Since initially \(v_C\) is zero, then constant value \(K_1\) is also zero, reducing \(v_C\) to

\[
v_C(t) = K_2 \sin(\omega_{LC} t)
\]

(A.5)

and, therefore, inductor current is

\[
i_L(t) = i_C(t) = C_{\text{MAX}} \frac{dv_C}{dt} = K_2 C_{\text{MAX}} \omega_{LC} \cos(\omega_{LC} t).
\]

(A.6)

Since the initial inductor current is the peak value \(I_{L,\text{PK}}\) from the preceding energizing step, \(K_2\) equals

\[
K_2 = \frac{I_{L,\text{PK}}}{C_{\text{MAX}} \omega_{LC}} = I_{L,\text{PK}} \sqrt{\frac{L}{C_{\text{MAX}}}}
\]

(A.7)

and the capacitor voltage and inductor current during the de-energizing step are

\[
v_C(t) = I_{L,\text{PK}} \sqrt{\frac{L}{C_{\text{MAX}}}} \sin(\omega_{LC} t)
\]

(A.8)

and
\[ i_L(t) = i_C(t) = I_{L,PK} \cos(\omega_{LC} t). \] \hfill (A.9)

Once the capacitor voltage reaches \( V_{BAT} \), \( S_{D1} \) and \( S_{D2} \) open to complete the precharge process.

### A.2 Two-Switch Precharger

The two-switch precharger in Figure A.1(b) also decomposes into an energize and de-energize sequence, except that during the first step the battery energizes both \( L \) and \( C_{VAR} \) when closing energizing switch \( S_E \). As shown in Figure A.3(a), during this time the circuit is a resonant LC circuit with the battery as a forcing input, which can be described with the following second-order differential equation

\[ \frac{d^2 v_C}{dt^2} + \frac{v_C}{LC_{MAX}} = \frac{V_{BAT}}{LC_{MAX}}, \] \hfill (A.10)

with the subsequent general solution

\[ v_C(t) = K_3 \cos(\omega_{LC} t) + K_4 \sin(\omega_{LC} t) + V_{BAT}, \] \hfill (A.11)

where \( K_3 \) and \( K_4 \) are constant values. Because initially \( v_C \) is ideally zero, constant \( K_3 \) is

\[ K_3 = -V_{BAT}, \] \hfill (A.12)

and the capacitor voltage is

\[ v_C(t) = V_{BAT} [1 - \cos(\omega_{LC} t)] + K_4 \sin(\omega_{LC} t). \] \hfill (A.13)

Therefore, the inductor current is

\[ i_L(t) = i_C(t) = C_{MAX} \frac{dv_C}{dt} = C_{MAX} \omega_{LC} [V_{BAT} \sin(\omega_{LC} t) + K_4 \cos(\omega_{LC} t)]. \] \hfill (A.14)

Since the initial inductor current is zero, then constant \( K_4 \) is also zero, and, as a result, the capacitor voltage and inductor current during the energizing step are

\[ v_C(t) = V_{BAT} [1 - \cos(\omega_{LC} t)] \] \hfill (A.15)
After energizing L and \( C_{\text{VAR}} \) with sufficient investment energy, opening \( S_E \) and closing de-energizing switch \( S_D \), after a dead time where both are open, de-energizes L into \( C_{\text{VAR}} \). During this time the circuit is a resonant LC circuit identical to the de-energizing circuit in the four-switch precharger, as shown in Figure A.3(b). Equations A.2 and A.3 also describe and represent the circuit, but with different initial conditions. If \( v_C \) begins the de-energizing step at initial voltage \( V_I \), constant value \( K_1 \) is now \( K_1 = V_1 \),

\[
K_1 = V_1, \tag{A.17}
\]

and the capacitor voltage is

\[
v_C(t) = V_1 \cos(\omega_{LC} t) + K_2 \sin(\omega_{LC} t) \tag{A.18}
\]

Therefore, inductor current is

\[
i_L(t) = C_{\text{MAX}} \frac{dv_C}{dt} = C_{\text{MAX}} [K_2 \omega_{LC} \cos(\omega_{LC} t) - V_1 \omega_{LC} \sin(\omega_{LC} t)]. \tag{A.19}
\]

Since the initial inductor current is the peak value \( I_{L,PK} \) from the preceding energizing step, \( K_2 \) equals
\[ K_2 = \frac{I_{L,\text{PK}}}{C_{\text{MAX}}\omega_{LC}} = I_{L,\text{PK}} \sqrt{\frac{L}{C_{\text{MAX}}}} \quad (A.20) \]

and the capacitor voltage and inductor current during the de-energizing step are

\[ v_c(t) = V_1 \cos(\omega_{LC} t) + I_{L,\text{PK}} \sqrt{\frac{L}{C_{\text{MAX}}}} \sin(\omega_{LC} t) \quad (A.21) \]

and

\[ i_L(t) = I_{L,\text{PK}} \cos(\omega_{LC} t) - V_1 \sqrt{\frac{C_{\text{MAX}}}{L}} \sin(\omega_{LC} t). \quad (A.22) \]

Since energizing \(L\) and \(C_{\text{VAR}}\) with sufficient investment energy ideally corresponds to precharging \(C_{\text{VAR}}\) to \(0.5V_{\text{BAT}}\), the capacitor voltage and inductor current are

\[ v_c(t) = \frac{1}{2} V_{\text{BAT}} \cos(\omega_{LC} t) + I_{L,\text{PK}} \sqrt{\frac{L}{C_{\text{MAX}}}} \sin(\omega_{LC} t) \quad (A.23) \]

and

\[ i_L(t) = I_{L,\text{PK}} \cos(\omega_{LC} t) - \frac{1}{2} V_{\text{BAT}} \sqrt{\frac{C_{\text{MAX}}}{L}} \sin(\omega_{LC} t). \quad (A.24) \]

After \(C_{\text{VAR}}\) absorbs and exhausts \(L\)’s energy, \(v_c\) reaches \(V_{\text{BAT}}\) and \(S_D\) opens to end precharge.
APPENDIX B

INITIAL PROOF-OF-CONCEPT PROTOTYPE

In the following section, the initial proof-of-concept energy-harvesting prototype is presented. The system precharges, detects, and synchronizes to a voltage-constrained variable capacitor to verify experimentally that it is possible to harvest energy from vibrations. To this end, the prototyped IC implements a four-switch precharger with externally powered control circuits. In addition, a harvesting diode channels the harvesting current that a manually-turned trimmer capacitor generates to the battery source (to avoid harvest detection circuits).

B.1 System

Since inductors are quasi-lossless devices, to maximize energy gain, the four-switch inductor-based precharger in Figure B.1 transfers energy from battery $V_{\text{BAT}}$ to variable capacitor $C_{\text{VAR}}$. The battery energizes inductor $L$ by imposing $V_{\text{BAT}}$ across it with switches $S_{E1}$ and $S_{E2}$, consequently causing inductor current $i_L$ to increases linearly until sufficient energy is stored, at which point switches $S_{E1}$ and $S_{E2}$ open and $S_{D1}$ and $S_{D2}$ close and channel the stored energy to $C_{\text{VAR}}$ at maximum capacitance, just before the onset of the harvesting phase. To simplify control, diode $D_H$ is used to connect the variable capacitor to the battery. As capacitance decreases, battery $V_{\text{BAT}}$ clamps capacitor voltage $v_C$ via $D_H$ and the resulting harvesting current $i_{\text{HARV}}$ charges the battery.
The precharge control block illustrated in Figure B.1 senses when to precharge $C_{VAR}$ (at $C_{MAX}$) and applies the proper gate-drive signals for a predetermined energizing time ($t_E$). Instead of sensing capacitance directly, which might require ac currents or voltages, the precharger detects when capacitance starts to decrease by monitoring $v_C$.

Since $C_{VAR}$ is in an open circuit condition (charge constrained) during its reset phase, $v_C$ increases as soon as $C_{VAR}$ starts to decrease from its maximum $C_{MAX}$. At this point, if $v_C$ is less than $V_{BAT}$, precharge commences, and because the precharge time is on the order of nanoseconds and therefore substantially shorter than the vibration period, which is in milliseconds, capacitance is close to its maximum value.

The time $t_E$ is preset to energize the inductor with sufficient energy to subsequently charge the capacitor to $V_{BAT}$. As derived in Equation 5.6, the energizing time is

$$t_E = \sqrt{\frac{L}{C_{MAX}}} \quad (B.1)$$
which is independent of $V_{BAT}$. Consequently, even as $V_{BAT}$ changes (for example, a Li Ion spans 2.7-4.2 V across its state of charge), a constant $t_E$ transfers sufficient energy to $C_{VAR}$. In practice, however, $t_E$ is set slightly higher to offset the energy losses associated with the transfer.

The precharge process ends as soon as capacitor voltage $v_C$ equals or surpasses battery voltage $V_{BAT}$, when all switches turn off. Any excess energy in the inductor, if sufficient, returns to the battery through the harvesting diode by charging $C_{VAR}$ a diode voltage above $V_{BAT}$. Some of the excess inductor energy might not be sufficient to forward bias either the harvesting diode or any of the body diodes across the switches, and thus causes oscillations between $L$ and parasitic capacitances in the circuit, which eventually die down because of parasitic resistances present. It is best to minimize any extra energy, however, to reduce the losses associated with transferring it through the system.

**B.2 Circuits**

The proof-of-concept energy-harvesting system prototype was fabricated with AMI’s 1.5-μm CMOS process technology. The complete system is comprised of the harvesting diode $D_H$, switches $MP_{E1}$, $MN_{D1}$, $MN_{E2}$, transmission gate $MN_{D2}$-$MP_{D2}$, and the control electronics (gate drivers, logic circuits, a timer, and state detectors). Part of the control electronics were kept off chip for experimental flexibility and reach. To comprehend the fundamental limits of the system, and those of voltage-constrained energy harvesters of this genre in general, all precharger components, on- and off-chip, were powered from a separate supply ($V_{DD}$).
Transistors MP_{E1} and MN_{E2} turn on to energize inductor L and the stored energy in L is driven to C_{VAR} when MN_{D1} and transmission gate MP_{D2}-MN_{D2} conduct, after MP_{E1} and MN_{E2} are off. Complementary switches MP_{D2}-MN_{D2} realize switch S_{D2} because MP_{D2} alone lacks gate drive to conduct enough current when the capacitor is initially discharged. Individual on-chip buffers drive each power switch, except for MN_{D2}, which shares its driving signal with MN_{D1}. The harvesting diode connecting the capacitor to the battery is a diode-connected NPN transistor (AMI’s 1.5-μm CMOS technology offers vertical n-type BJTs).

To detect the state of C_{VAR}, and thus determine when to precharge it, an off-chip state detector circuit is used. Comparator CP_{V-DETECT} detects when v_{c} falls below V_{BAT}, which is the first condition required to start the precharge process. The propagation delay of this comparator should be sufficiently short to ensure the precharge phase stops before C_{VAR} charges above its target value (V_{BAT}), which could otherwise incur additional losses.
in the system. Slope-detecting comparator CP\textsubscript{SLOPE} identifies whether or not the second condition, that $v_C$ rises, is met by comparing $v_C$ with a delayed version of itself ($v_C\text{\_DELAY}$). That way, if $v_C$ increases (or decreases), $v_C$ is lower (or higher) and CP\textsubscript{SLOPE} therefore asserts an enabling (or disabling) signal. A 5-MΩ and 4.7-nF RC circuit implements a delay of approximately 20 ms and buffer OP\textsubscript{BUFFER} isolates $C_{VAR}$ from the RC circuit. The comparators include some hysteresis to desensitize the circuit to glitches and any other extraneous noise present.

When the aforementioned conditions are met, on-chip logic determines the appropriate precharge switching sequence, including dead time between oppositely phased digital signals to avoid transient shoot-through (short-circuit) power losses in the precharger switches. Logic gate AND\textsubscript{1} enables a timer with signal $v_{\text{START}}$ and activates the energizing process of inductor $L$ via signal $v_{\text{ENEG}}$, while $v_{D\text{-ENEG}}$ is low. After the timer reaches its preset value, it flags the logic to stop energizing $L$ with signal $v_{\text{STOP}}$, which forces $v_{\text{ENEG}}$ to drop and, after a dead-time delay (where all switches are off), prompts the circuit to de-energize $L$ via $v_{D\text{-ENEG}}$, when it goes high. The combined propagation delay of logic gates AND\textsubscript{2}, AND\textsubscript{3}, and NOR determines the duration of dead time. The de-energizing switches remain on until $v_C$ is again greater than $V_{\text{BAT}}$, at which point the logic disables the timer and shuts off all MOS switches.

The energizing time of the inductor is set with the timer circuit shown in Figure B.3. Once reset and enabled by the logic block (i.e., $M_{\text{RESET}}$ is turned off), the circuit triggers a cut-off signal when comparator CP\textsubscript{TIMER} senses that linearly increasing ramp voltage $v_{\text{RAMP}}$ surpasses preset reference voltage $V_{\text{TIME-REF}}$, the latter of which effectively sets the total energizing time for the inductor. Charging on-chip capacitor $C_{\text{RAMP}}$ with a
A constant current-source reference produces $v_{\text{RAMP}}$. The current source is realized by forcing reference voltage $V_{\text{I-REF}}$ across resistance $R_{\text{I-REF}}$ via the negative feedback loop comprised of op-amp $\text{OP}_{\text{I-REF}}$ and transistor $\text{MN}_{\text{I-REF}}$. This reference current is subsequently mirrored and amplified by a factor of five (i.e., $M = 5$) before channeling it to $C_{\text{RAMP}}$, resulting in

$$I_{\text{RAMP}} = M \frac{V_{\text{I-REF}}}{R_{\text{I-REF}}}, \quad \text{(B.2)}$$

which linearly charges $C_{\text{RAMP}}$ with a slope of

$$\left( \frac{\Delta V}{\Delta t} \right)_{\text{RAMP}} = \frac{I_{\text{RAMP}}}{C_{\text{RAMP}}} \quad \text{(B.3)}$$

and defines $V_{\text{TIME-REF}}$ to

$$V_{\text{TIME-REF}} = t_{\text{E}} \left( \frac{\Delta V}{\Delta t} \right)_{\text{RAMP}}. \quad \text{(B.4)}$$

In practice, parasitic capacitors in parallel to $C_{\text{RAMP}}$ slow the rising ramp rate so $V_{\text{TIME-REF}}$ should be lower.
B.3 Experimental Results

The power switches, gate drivers, digital logic, diode, and part of the timer were fabricated on 1.4 x 1.8 mm$^2$ of the 2.2 x 2.2 mm$^2$ die shown in Figure B.4 using AMI’s 1.5-µm CMOS process. A 3-V supply emulated a moderately charged Li Ion ($V_{BAT}$), whose full range normally spans 2.7 to 4.2 V. The off-chip surface-mount 4 x 4 x 2 mm$^3$ inductor package used had a measured inductance of roughly 10.72 µH with an ESR of 240 mΩ. Manually turning a trimmer capacitor with a measured maximum-minimum capacitance range of approximately 250 to 60 pF (including parasitic capacitances present) emulated the harvesting device under vibration conditions. Although continually turning the manual capacitor to charge a battery would have been ideal, the process was impractical because of its non-periodic nature and the human element of fatigue; however, the objective of the set-up was to test the viability of the harvesting scheme on a per cycle basis, not its steady-state behavior.

Figure B.4. Die photograph of the 2.2 x 2.2 mm$^2$ proof-of-concept prototype IC.
To prove harvesting is possible by constraining voltage, the capacitor is manually precharged and turned. By momentarily shorting the variable capacitor to the supply after setting it to its maximum capacitance, manually precharges the device. The capacitor is then turned and its resulting current recorded. Turning the capacitor, however, is a manual process and the device’s response is consequently nonlinear, introducing what

Figure B.5. (a)-(b) Sample harvest measurements by directly precharging variable capacitor $C_{\text{VAR}}$.

To prove harvesting is possible by constraining voltage, the capacitor is manually precharged and turned. By momentarily shorting the variable capacitor to the supply after setting it to its maximum capacitance, manually precharges the device. The capacitor is then turned and its resulting current recorded. Turning the capacitor, however, is a manual process and the device’s response is consequently nonlinear, introducing what
appears to be noise. The finite delay between the initialization process and actually turning the device further introduces inaccuracies, allowing parasitic leakage currents to discharge the capacitor slightly from its initial value. Nonetheless, Figure B.5 shows the harvesting currents the variable capacitor drive through the diode to the battery supply for two different sets of measurements. Integrating the power harvested, which is the product of the measured current and battery voltage $V_{\text{BAT}}$, over the cycle time yielded 6.11 and 6.37 nJ for the results shown and an average of 5.82 nJ across eight separate measurements.

The precharger was then tested by allowing the system to detect the conditions necessary for a precharge cycle, initiate the sequence, and charge variable capacitor $C_{\text{VAR}}$ to battery voltage $V_{\text{BAT}}$. To this end, $C_{\text{VAR}}$ was set at its maximum capacitance point and turned. Figure B.6 shows how both conditions for precharge are detected: $v_{\text{COND1}}$ is high when $v_C$ is less than or equal to $V_{\text{BAT}}$ and $v_{\text{COND2}}$ is high when $v_C$ increases (i.e., when $v_C$ is greater than its delayed version $v_{C\text{-DELAY}}$). Figure B.7(a) shows control signals $v_{\text{ENEG}}$ and $v_{\text{D-ENEG}}$ when subjected to this test, the former of which instructs the system to energize inductor $L$ and the latter to release its stored energy to $C_{\text{VAR}}$. As $L$ is energized, $V_{\text{BAT}}$ (3 V) is impressed across $L$ (inductor voltage $v_L$ is shown in Figure B.7(b)), forcing an inductor current of 17.61 mA (on average) and resulting in an average invested energy per cycle of 1.66 nJ. When de-energizing $L$, $v_L$ is reversed by connecting $L$ to $C_{\text{VAR}}$, gradually releasing energy to $C_{\text{VAR}}$ and consequently increasing its voltage $v_C$. Once precharge ends, $L$ retains some remnant excess energy, which causes it to resonate with the surrounding parasitic capacitances. Parasitic resistances eventually dampen the oscillations and the inductor voltage returns to zero volts.
After determining the precharger was functional, full system-level experiments were performed, verifying precharge and harvest automatically cycled with variations in $C_{VAR}$, as designed and shown experimentally in Figure B.8. The average energy per cycle directed to the battery over 126 different sets of measurements was 9.7 nJ/cycle (Figure B.8(a) is a sample run), giving a net energy gain (by subtracting the investment from the harvest) of approximately 8 nJ/cycle. Figure B.8(b) illustrates the results of continually

Figure B.6. (a) Variable capacitor voltage $v_C$ and its delayed version $v_{C\text{-DELAY}}$ and (b) corresponding state detector outputs $v_{COND1}$ and $v_{COND2}$. 
turning the trim capacitor for six consecutive cycles, harvesting a total of 63 nJ over the span of the six cycles shown (without subtracting the corresponding investment energy).

B.4 Conclusion

Fundamentally, the experiments successfully confirmed that it is possible to harvest kinetic energy with a voltage-constrained variable capacitor. An important feature
of the proposed solution is its ability to automatically detect when to precharge the capacitor without having to sense or measure capacitance directly, or accurately pre-synchronize the system to a predetermined frequency of capacitor variations. However, the prototype did not integrate the entire harvester-charger circuit within a single IC and, furthermore, the battery did not power the entire system.
REFERENCES


VITA

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