Powering Low-Cost Utility Sensors using Energy Harvesting

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Keywords

«Power management», «wireless sensors», «contactless energy transfer», «supercapacitor»

Abstract

This paper presents the use of magnetic-field and solar energy harvesting to power low-cost utility sensors. As we move towards a more dynamic and intelligent electricity grid, inexpensive smart sensors that help in extracting information on current, temperature and voltage from different utility assets are becoming vital for maintaining high reliability of the power grid. Maintenance free sensors that do not require batteries for operation can be developed with the use of efficient energy harvesting techniques. This paper presents a novel 0.2 V to 3.3 V AC-DC boost converter which converts magnetic field (H-field) and solar energy to electrical energy used by the sensor electronics. A supercapacitor has been used to operate the sensor even in an outage. Numerous design constraints have been identified. Further, a low-cost power circuit has been fabricated and experimentally tested to show functionality under varying operating conditions. Finally, analysis under faulted conditions has also been presented to show the robustness of the power circuit.

Introduction

Grid assets, such as conductors, transformers, cables, shunt capacitors etc., presently used by utilities are old and are being stressed due to the rapid increase in power demand. To increase reliability of the power grid, effective asset utilization, and perform condition based asset maintenance, monitoring of critical asset parameters such as current, voltage and temperature, is essential [1]. This requires installation of sensors on all the grid assets. However, most of the sensors presently available in the utility domain are very expensive, large and bulky. They require clamping around the asset and are limited by their geometry [1]. In addition, some of the sensors also need batteries for operation. As batteries need replacement after every few months, the overall maintenance expenditure of the sensor is increased. The present state-of-the-art sensors have prices ranging in the order of thousands of dollars and thus, implementation of even 100 sensors in a substation requires initial investments on the order of half a million dollars only on sensors (assuming the cost of the sensors to be $ 5000). This investment is exorbitantly high to be justified for mass implementation of present day sensors on utility assets. Evidently, cost of present day sensors need to be reduced by an order of magnitude if they were to make a strong business case.

As compared to the Business as usual (BAU) scenario, a new idea was introduced in [2] to use a universal distributed sensing solution which is not limited by asset geometry and can be used for
monitoring different kinds of assets. The nodes in such a distributed sensing solution are connected by a wireless sensor network (WSN) architecture. Consequently, realizing benefits of low power protocols, self healing networks, increased security, meshed multi-hop topologies, increased coverage area and all in all, a highly reliable communication network for relaying information. This research focused at developing a wireless sensor which uses an open core coil arrangement for measuring current. An open core configuration allows the sensor to be either kept in the vicinity of or stuck on to a utility asset for normal operation and therefore, removes the constraint of clamping around the utility asset for monitoring current. These sensors are referred to as Stick-on sensors. The developed sensors can be used in conjunction with a variety of grid assets and are not limited to monitoring only transmission / distribution lines or cables. Moreover, the size and cost of the sensor can be reduced considerably as compared to conventional solutions. Nonetheless, as compared to a conventionally used closed core solution it is quite challenging to harvest energy from the magnetic field (H-field) with a small open core. This paper introduces a novel power management architecture which tackles the challenges related to self powering these small low-cost sensors. Solar energy harvesting has also been used in addition to H-field energy to increase reliability of operation of the sensor. Moreover, operation of the power management circuit in faulted modes is also considered and solutions have been provided in this paper.

**Energy harvesting for powering smart sensors**

Typical power requirements of the targeted utility sensors are on the order of 10s of mW during sensing, 100s of mW during transmission of sensed data, and 100s of µW during sleep mode. These sensors are operated with a very low duty cycle, typically once every 15-30 min and therefore, their average power requirement is very low. Nonetheless, a typical sensor requiring 25 mA active mode current for 1 sec, 100 µA sleep mode current for 10 min and a 3.7 V, 1200 mAh battery would last for not more than a year. Therefore, relying on batteries for powering these sensors is clearly not a feasible solution for implementing a distributed sensor network on the power grid.

Battery dependence of the utility sensors can be removed by harvesting energy from the ambient. Two major sources of energy for powering utility sensors are electric and H-fields which are present in abundance near most of the utility assets. With the use of an x-shaped open core (XFC) depicted in Fig. 1, it is possible to harvest hundreds of mWs from the H-field when sufficient current is flowing in the utility asset (90 mW at 800A conductor current). However, in the case of electric field energy harvesting, the size of the harvester needs to be increased significantly to harvest similar amount of power, 95 mW for a 225 cm² plate capacitor [3]. Therefore, the energy density attained from H-field is considerably high as compared to electric field. In addition, due to high voltages involved with electric field energy harvesting the feasibility of implementation decreases and cost increases. In essence, H-field emerges as the clear winner over electric field for harvesting energy.

![Magnetic field EH](image1.png)  
**Thickness = 18mm, Winding Turns = 300**  
*a) Magnetic field EH*  

![XFC stuck-on to a utility conductor](image2.png)  
**b) XFC stuck-on to a utility conductor**  

Fig. 1: Magnetic field energy harvester (EH)

Nevertheless, due to the small size and open core configuration of a Stick-on utility sensor, it faces a major challenge related to self powering. A plot of variation in maximum harvested power at different primary current values and different distances of the XFC from the conductor is shown in Fig. 2. It can be seen that as the primary current magnitude reduces or as the distance of the XFC from the conductor increases, the maximum energy harvested by the XFC goes down exponentially. For instance, when the XFC is kept at a distance of 9 inches from the conductor it can harvest no more
than 5 mW even in the presence of 900A of conductor current. Therefore, it becomes critical to efficiently convert the available magnetic field energy to electrical energy at these operating conditions. Furthermore, the power circuit has to be designed in a way that ensures low power consumption, low component count, minimal active circuitry, and effective use of sleep modes.

Boost converter for low voltage energy harvesting

When there is insufficient energy in the magnetic field, low voltages are induced on the energy harvester (EH) terminals. The XFC has an open circuit voltage \( V_{oc} \) of nearly 200 mV at 60 A of conductor current; this \( V_{oc} \) is not even sufficient to overcome the forward threshold of semi-conductor switches and diodes. As the electronic circuits on the sensor board require regulated DC voltage supply (typically 2 – 3.3 V), high voltage boost functionality is required.

Conventional approaches

One approach of boosting voltages is through the use of charge pumps such as the traditional Cockcroft Walton charge pumps as shown in Fig. 3a. Although, the output voltage goes up with an increase in number of stages, but it necessitates series connection of multiple diodes which increases losses and reduces effective output voltage due to aggregated diode drop. In addition, if the parasitic capacitances \( C_p \) become comparable to coupling capacitors, the voltage boost is limited to twice the input voltage irrespective of the number of stages of the charge pump. Numerous topologies for effective charge pumps have been proposed in the past but all suffer from one or the other disadvantage discussed above [4]. A recent research presented in [5] developed a hybrid configuration that used an inductive boost converter followed by a two stage charge pump. This circuit could boost voltages from 0.2V to 1.2 V (still not 3.3 V DC required by sensor electronics) for a thermopile application. Some other research efforts have tried to analyze and implement charge pumps in other energy harvesting applications, such as solar and RFID, but do not look promising in utility applications.

Others have used simple AC/DC bridge rectifiers followed by a DC/DC boost converter to realize the boost functionality, shown in Fig. 3b [6]. However, this circuit is not an effective solution for energy harvesting applications as it has- two energy conversion stages which decreases overall efficiency; increased device count which increases cost; and low input voltages which may not be high enough to overcome the forward threshold voltages of semiconductor devices.

Fig. 2: Variation of maximum harvestable power with change in distance of the XFC from the conductor and change in conductor current

Fig. 3: Conventional boost converters
The main idea is to reduce the number of energy conversion stages so that the efficiency of the circuit can be improved without compromising the voltage boost functionality. One simple solution to this problem is shown in Fig. 4a [7]. This converter allows direct AC/DC boost conversion and reduction of the number of energy conversion stages to only one. However, it requires two separate inductors for operation, increased component count, and the need for isolated gate drive circuits for practical implementation. In addition, this converter requires the knowledge of voltage polarity of the supply for proper operation (as switches S1 and S2 are pulsed alternately in positive and negative half cycles). A complex control strategy results in the use of additional active circuit components and increase in power consumption. Recently, based on this idea, direct AC-DC boost converters for micro-generators have been proposed [8]. However, almost all the existing topologies require batteries, multiple diodes and capacitor combination for start-up, and therefore they cannot be used for utility applications.

**Proposed approach**

This research proposes a simple and a novel technique of realizing boost functionality in energy harvesting applications. The proposed converter is shown in Fig. 4b. This circuit uses a bi-directional switch connected in a common source configuration. The switches can be pulsed together and do not require a floating gate circuit as the ground of the circuit is common with the source of the MOSFETs. Moreover, the circuit uses the leakage inductance of the EH and transformer as the energy transfer element and does not require an external inductor. Voltage boost operation is obtained by using the AC switch that charges the inductor whenever it is ON. While, when the AC switch is turned OFF, diode D1 (or D2 depending on the current direction) conducts and the energy is transferred from the leakage inductor to the load. The operation of the converter is depicted in Fig. 5.

![Diagram of Proposed Converter](image-url)

**Fig. 4:** a) A Direct AC-DC boost converters for energy harvesting applications, b) proposed 0.2 V to 3.3 V AC/DC boost converter with a wide operating range

![Diagrams of Modes of Operation](image-url)

**Fig. 5:** Modes of operation of the proposed converter
The proposed converter can be operated in either the continuous conduction mode (CCM) or the discontinuous conduction mode (DCM). The conceptual waveforms that are realized in both these modes are given in Fig. 6. In CCM, the current through the inductor is time varying at the frequency of supply with a ripple at the switching frequency. However, the converter will not be able to operate in pure CCM at all times as at lower inductor currents the converter is bound to enter DCM operation. Therefore, it is difficult to compute a closed form expression for the output voltage. Nonetheless, an approximate CCM mode output voltage can be derived using a large signal (RMS) approximation of the circuit given in (1). While, in the case of pure DCM, the inductor current goes to zero after every switching period and therefore volt-second balance can be applied in one switching period. Again using a large signal model the output voltage in DCM is given by (1).

\[
V_{o,CCM} = \frac{nV_{in}}{\sqrt{2}} \left( \frac{1-D}{R_L/R + (1-D)^2} \right)
\]
\[
V_{o,DCM} = \frac{nV_{in}}{2\sqrt{2}} \left( 1 + \sqrt{1 + \frac{2D^2R}{f_s/Ln^2}} \right) \tag{1}
\]

Where \( n \) is the transformer turns ratio, \( V_{in} \) is the peak input voltage, \( R \) is the load resistance and \( R_L \) is the inductor resistance.

Combination of computed analytical voltage boost expression with the simulation results is shown in Fig. 7a. The simulation results show that the converter is able to boost voltages as low as 0.2 V to output voltages much greater than 3 V. However, large output voltages are achieved at the cost of higher current in switches as the inductance is relatively small. In a practical application, the harvested power is limited. Therefore, as the load current increases, the input voltage reduces to keep power limited to its maximum value which, consequently, reduces the output voltage (a phenomenon not observed if the voltage source is stiff).

**Simulation Results**

Simulation of the proposed converter was performed in Synopsys Saber™ to validate the voltage boost functionality. The results of the simulations are presented below.

**DCM Operation**

The circuit parameters in the simulation were chosen to be, \( V_F = 0.2 \) V peak, frequency = 2 kHz, \( L_F = 10 \) µH, Transformer step up ratio = 1:20, \( C = 10 \) µF and \( R = 10 \) kΩ, MOSFET body diode drops = 0.7 V and Schottky diode forward drop = 0.3 V (see Fig. 4b to refer the variable names). The conformance of computed analytical voltage boost expression with the simulation results is shown in Fig. 7a. The simulation results show that the converter is able to boost voltages as low as 0.2 V to output voltages much greater than 3 V. However, large output voltages are achieved at the cost of higher current in switches as the inductance is relatively small. In a practical application, the harvested power is limited. Therefore, as the load current increases, the input voltage reduces to keep power limited to its maximum value which, consequently, reduces the output voltage (a phenomenon not observed if the voltage source is stiff).

**Combined CCM and DCM Operation**

Same circuit parameters were chosen as before with a change in the value of inductance to 0.5 mH. A plot of output voltage obtained from simulation results versus analytical results is shown in Fig. 7b. The approximate analytical results follow the simulation results at lower duty cycle but fail to follow at higher duty ratios. The output voltage of the converter decreases at higher duty ratios due to the small equivalent series resistance (ESR) of 1 mΩ included with the inductor.
In practical applications, the transformer and XFC will have a larger ESR which can dramatically change the output voltage characteristics. Output voltages obtained at two different values of ESRs (1 \( \text{m\,\Omega} \) and 200 \( \text{m\,\Omega} \)) are compared in Fig. 7c. It can be seen that the converter has characteristics very similar to a DC/DC converter. Thus, it is advisable to operate the converter at a duty cycle less than 80% to obtain maximum voltage boost.

![Fig. 7: Comparison of output voltage obtained using simulation results versus analytical results in a) pure DCM, and b) combined CCM and DCM, and c) comparison of change in output voltage with duty cycle at two different ESRs](image)

**Power circuit design and experimental results**

A prototype was constructed to validate the operation of the proposed converter in practice, as depicted in Fig. 8a. Another prototype was also built using mainly surface mount components to prove that the size of the circuit can be reduced even further, as shown in Fig. 8b. The power circuit was used to operate a wireless Stick-on utility sensor for current and temperature sensing of a conductor and hence the circuit also comprises extra signal conditioning components. The power circuit used two 35 V N-channel MOSFETs, two low-forward drop Schottky diodes and a 1:20 step-up transformer. The results of the experiment are shown in Table 1. The efficiency of the converter was calculated to be 75% at 200A conductor current. The current and voltage waveforms at different points in the circuit are shown in Fig. 9a and 9b. Further, Table 1 shows voltage boost obtained at different duty cycles.

![Fig. 8: a) Proposed converter circuit board, b) a comparison of size between through-hole and surface mount prototype, and c) the power circuit was used to operate a TI CC2530 Zigbee wireless sensor](image)

**Table I: Converter operation at different duty cycles when primary current is 60 A, load resistance is 50 k\( \Omega \) and switching frequency is 2 kHz**

<table>
<thead>
<tr>
<th>AC Voltage (RMS Volts)</th>
<th>Duty Cycle (%)</th>
<th>DC Voltage (Volts)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.2</td>
<td>30</td>
<td>3</td>
</tr>
<tr>
<td>0.2</td>
<td>50</td>
<td>3.5</td>
</tr>
<tr>
<td>0.2</td>
<td>70</td>
<td>4.4</td>
</tr>
</tbody>
</table>

**Black-Start Functionality**

Powering the sensors after an outage is a major challenge as the capacitor might have discharged completely. Thus, in the absence of a DC supply, gating pulses are not generated and the voltage boost functionality is not realized. The issue of blackstart was tackled by using a CMOS astable
multivibrator which bootstraps the output voltage using a positive feedback. Consequently, the power circuit was able to self start at primary currents as low as 60 A. As observed in Fig. 9c, when the primary current was increased from 15 A to 60 A, the DC voltage built up from 0.4 Vdc to 3.3 Vdc.

![Figure 9: The converter is operated with a 50% duty cycle, a) and b) show screenshots for $V_F$ (0.2 V/Div.), $V_T$ (2 V/Div.), $I_F$ (10 mA/Div.), $I_T$ (2 mA/Div.), $I_{D1}$ and $I_{D2}$ (1 mA/Div.), and c) Autonomous powering at 60A primary current, $V_o$ (1V/Div.), $V_F$ (0.2 V/Div.), Time base (2.5 Sec/Div.)](image)

**Operating range**
Robust operation of the power circuit over 60-1000 A has been demonstrated, as shown in Fig. 10a. At higher primary currents the input voltage can increase and surpass the reverse blocking voltage ratings of the semiconductor devices. A zener diode was used at the output side to clamp the output voltage to be within the safe operating area (SOA) of all the devices.

**Operation in an outage**
In an outage, in most utility applications the sensors are expected to operate only once to inform the coordinator about the loss of power in the asset. This reduces the on-board energy requirements enabling the use of maintenance free super-capacitors as backup. The designed circuit uses a 1 F supercapacitor (supercap). The circuit uses another 100 μF capacitor for handling power impulses which cannot be supplied by the supercap and supporting the DC bus voltage when the supercap is discharged. In the designed circuit for the sensor shown in Fig. 8c, a fully charged supercap can support more than 10 data transmission cycles in an outage over a period of nearly 2 hrs.

Further, the supercap cannot be connected directly across the DC bus of the power circuit as it poses as a low impedance load in the uncharged state. Therefore, the supercap was connected through a constant current source. Overcharging was avoided by clamping it to the bus voltage through diode D3. The supercap charging circuit is shown in Fig. 10b. The capacitor charging equation is given as

$$C \frac{dV_c}{dt} = \left( \frac{V_{D1} + V_{D2} - V_{BE}}{R_E} \right) = I_E$$

(2)

![Figure 10: a) Circuit operating at 1000A primary current, b) Supercap charging circuit](image)
Solar Energy Harvester Design

Most utility assets are present in open areas with abundance of sunlight. Hence, solar energy can act as another energy source for the sensor. However, solar energy has a diurnal variation and cannot be used as a primary source. It can, however, be used as a source of trickle charge for the supercap. In this way, the rating, size and cost of the solar cell can be kept very low, and minimal dependence on solar power can be realized.

In the proposed circuit, the solar cell does not necessarily operate at the maximum power point at all times, yet it is important to select a solar cell which has a high fill factor (FF) and thus a high efficiency. A simple rule of thumb to ensure high FF is to select a solar cell with a higher $V_{oc}$. In the proposed circuit, a solar cell having a $V_{oc}$ of 4 V and $I_{sc}$ (short circuit current) of 2.5 mA at 1000 W/m$^2$ insolation, shown in Fig. 11a was used. The cell was composed of eight cells in parallel each having a $V_{oc}$ of 0.5 V. The current density of the considered solar cell was equal to 34.6 mA/cm$^2$. The maximum harvestable power of the solar cell was characterized in the laboratory for different operating conditions and the results are presented in Table 2. The supercap was connected directly to the solar cell through a Schottky diode ($D_4$), as shown in Fig. 10b.

**Table 2: Characterization Results of the Solar Cell at 25 °C**

<table>
<thead>
<tr>
<th>Insolation (W/m$^2$)</th>
<th>$I_{sc}$ (mA)</th>
<th>$V_{oc}$ (V)</th>
<th>FF</th>
<th>Power at MPP (mW)</th>
<th>Efficiency at MPP (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>400</td>
<td>0.74</td>
<td>3.97</td>
<td>0.8</td>
<td>2.35</td>
<td>10.2</td>
</tr>
<tr>
<td>800</td>
<td>1.8</td>
<td>4.07</td>
<td>0.81</td>
<td>5.93</td>
<td>12.8</td>
</tr>
<tr>
<td>1000</td>
<td>2.5</td>
<td>4.08</td>
<td>0.81</td>
<td>8.26</td>
<td>14.3</td>
</tr>
</tbody>
</table>

To compute the time taken to fully charge the capacitor after an outage when a solar cell is used by the sensor, a piecewise linear characteristic for the solar cell can be considered. The linear characteristic is chosen such that the FF is kept constant at 0.81 (see Table 2) by fixing the maximum power point at 90% of $I_{sc}$ and 90% of $V_{oc}$.

From Fig. 11b, it is clear that the active mode occurs for 0.4 sec. Moreover, 15 mA and 120 µA of average load and quiescent current was consumed respectively. Consider that the supercap is completely charged ($V_c$ is 3.3V) when an outage occurs. During the active mode, based on the discharge equation (3), the voltage of the supercap will reduce. After the active mode the sensor goes into sleep mode and charging of the supercap begins. The time taken to charge the supercap back to 3.3 V will determine the time when another packet of data can be sent by the sensor. This would also be the minimum operating time between any two sensor operations. This minimum charge time of the supercap can be computed and is given by $T_c$ (3).

$$C \frac{dV_c}{dt} = \begin{cases} I_{sc}(V_c) - I_L, & \text{Active mode} \\ I_{so}(V_c) - I_Q, & \text{Sleep mode} \end{cases}$$

$$T_c = \frac{C}{M} \ln \left( \frac{(K_2 - K_1)/M + (K_1/M + 3.3)e^{-0.4M/C}}{K_2/M - 3.3} \right)$$

(3)

where $K_1 = I_{sc} (1 - V_{ds}/(9N V_{oc})) - I_L$, $K_2 = K_1 + I_L - I_Q$, $M = I_{so}/(9N V_{oc})$, $I_{so}(V_c)$ represents the linearized characteristic of the solar cell, $I_L$ and $I_Q$ are the load and quiescent current demanded by the power circuit, and $N$ is the number of parallel connected solar cells (in this case eight).

Another scenario that should also be considered is the time taken to start the sensor from a zero charge state (blackstart). In this particular case, the supercap is required to be charged from 0 V to 2 V (2 V is sufficient for powering all the analog circuitry and microcontroller). Fig. 11c shows a plot of the minimum charge time of the supercap and the blackstart time.

It can be seen from Fig. 11c that even at a low insolation level, 100 W/m$^2$ in the case of a cloudy day, the minimum charge time of the supercap is close to 100 sec (~2 min). In other words, in the presence of a solar cell, in an outage the sensor can be operated every 2 min, sufficiently low for most utility assets. Further, the time taken to start the sensor from a completely discharged state ranges from 14 min in high insolation levels (at 1000 W/m$^2$) to 45 min in average insolation levels (at 500 W/m$^2$),
again suitable for most applications. Furthermore, in normal operation, the rate of operation of the sensor is increased even further due to the presence of an additional current from the source side. This analysis clearly shows that with the addition of an inexpensive solar cell which trickle charges the ultra capacitor, a utility sensor becomes an even more attractive solution.

![Solar Cell Diagram](image)

**Fig. 11:** a) Eight parallel connected solar cells, b) current consumption of the fabricated sensor in the different modes of operation and c) supercap recharge time curve in an outage using a solar cell

### Operation in Faulted Conditions

Normal operation of the power circuit over a range of 60A – 1000 A has been tested in the laboratory. The open core configuration provides a highly linear input/output characteristic and the core does not saturate over a large operating range. However, in practical scenarios under faulted conditions, the current in the primary side can increase up to 20 kA or even higher for a few cycles, causing an increase in the V_{oc} and I_{sc} of the EH, and saturation of the core. As the core does not form a closed loop, only minor saturation effects are seen in the form of distortion in V_{oc} as shown in Fig. 12a. Further, given that a small short circuit resistance is used for measuring current, the I_{sc} waveform remains sinusoidal under symmetrical faults, and can still be used to measure current. However, at 20 kA, the magnitude of V_{oc} and I_{sc} may become as high as 40 V and 16 A respectively, well above the SOA of the operating devices. Further, as seen from (4) that high di/dt on the primary current also tend to increase the induced voltages on the windings. Therefore, the power circuit needs to be protected under faulted conditions.

\[ V = -\mu_0 \mu_n \frac{n A}{l} \frac{dI}{dt} \Rightarrow V = \frac{n \mu_0 \mu_n A o l I}{l} \cos(\omega t) \]  

The power circuit can be protected by using back-to-back zener diodes, bidirectional silicon transient voltage suppressor (TVS) diodes or metal oxide varistors (MOV) on the winding terminals of the EH. All these techniques clamp the short-duration high voltage induced on the windings in faulted conditions. This way the voltage is limited within the SOA of semiconductor devices and the excessive energy is dissipated in the protection devices. The bidirectional TVS diodes and MOVs both provide promising characteristics for suppression of high voltage surges and have the capacity of handling very high surge currents over multiple cycles as compared to zener diodes. Moreover, TVS diodes having fairly low breakdown voltages (V_{br}) on the order of 2 - 3 V can be found as opposed to MOVs which have relatively higher V_{br} (>10 V). Further, the clamp voltage of an MOV increases with an increase in current, whereas the TVS diodes have a relatively stiff clamp voltage. On the flip side, TVS diodes have a lower surge current rating and power dissipation capacity as compared to MOVs; consequently, a single MOV can replace multiple parallel connected TVS diodes. Both the protection devices have their own trade-offs and can prove to be useful in different applications.

In the case of the proposed energy harvesting circuit, if 3V TVS diodes are connected directly at the EH terminal, the voltage at the primary of the transformer will be clamped at ±3V. Given that a 1:20 transformer is used, the secondary of the transformer will experience ±60 V stress, well above the SOA of the switching devices. This would require the need of a clamping zener diode on the DC side, to keep the voltages within SOA of the devices. However, the zener will need to clamp voltages from
60V to as low as 6 V and would be required to handle high power dissipation, thereby increasing its size. Alternatively, the protection devices can be connected on the secondary of the transformer, this way they can have a higher leeway on clamp voltage allowing the use of MOVs.

In the case of the Stick-on sensor, by way of an example, consider a primary faulted current of 20 kA which persists for 5 cycles. A 10 V bidirectional TVS diode (having a 15 V clamp voltage) which can support a peak surge current of 97 A with a safe energy dissipation of nearly 3.6 J (SMCJ9.0 by Bourns) can be used to protect the power circuit. If it was assumed that all the short circuit current flows through this TVS diode, the loss after 5 cycles is approximately equal to 15 J which can be handled by using 5 of these diodes in parallel. As shown in Fig. 12b, an MOV can also be connected in parallel as it has a higher surge energy capacity and can handle excessive surges, for instance, during lightning strikes.

![Fig. 12. a) Isc and Voc waveforms at 10 kA primary fault current obtained from FEM simulation using an I-shaped steel 1010 core in Maxwell 3D™, b) protection architecture for the proposed circuit](image)

**Conclusion**

This paper presented novel circuit design techniques to power utility sensors using magnetic-field and solar energy harvesting. Solar energy was used as a source to trickle charge the supercap for operation in an outage and to support operation when there is low energy in the magnetic field. A novel 0.2 V to 3.3 V boost converter was designed and tested through extensive simulation studies and experiments. Finally, issues related to operating the power circuit under faulted conditions were analyzed and solutions were provided. The power circuit proposed in this paper is a robust solution for powering smart utility sensors.

**References**


