

Power Management Electronics

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Introduction

In most low-power systems, power management is generally thought of as being an ability to switch certain parts of a system off or put them in a low power state when they are not required, and to manage the charging of a battery. Whilst these are important aspects of low power electronics powered by energy harvesters, there are much more fundamental reasons for requiring power electronics in an energy harvesting system than simply managing a battery and conserving energy:

- In order to achieve high power density from the energy harvester, there should be some form of impedance match between the energy source and transducer and the electrical system. This requires control of the input impedance of the circuit which interfaces to the transducer
- The output voltage and current from the energy harvester are rarely directly compatible with load electronics and thus some form of voltage regulation is required
- As discussed in Chapter 3, some form of energy storage is almost certainly required so that the intermittency of the energy harvesting source does not have a detrimental effect on the continuous operation of the system

Therefore, the basic power electronics topology for an energy harvesting system often follows that shown in Figure 1.

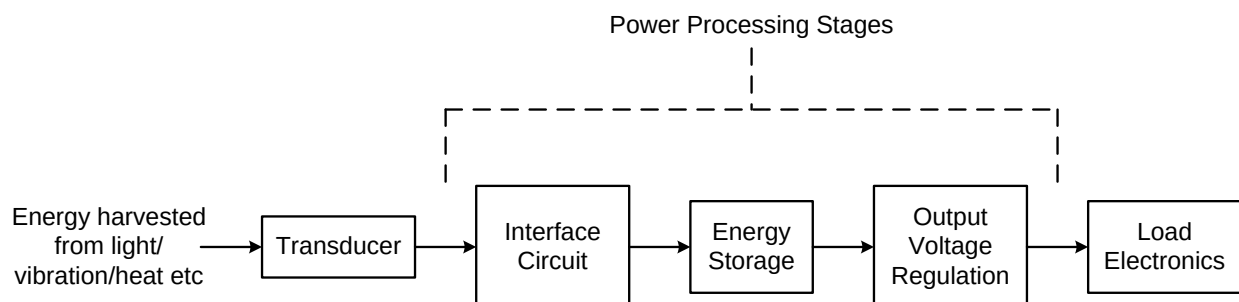


Figure 1 Power Electronics Topology for Energy Harvesting Systems

Interface Circuit Impedance Matching

In a large scale electrical energy generation plant such as a coal fired power station, where large amounts of power are produced and where fuel must be purchased, it is important that as much of the energy contained in the original fuel source as possible is converted into useful electrical power. This

firstly requires a high efficiency of conversion of the energy stored in the fuel to a mechanical form, secondly a high conversion efficiency of that mechanical energy to electrical energy and finally a high efficiency of power transfer from the electrical generator to a load. In order to ensure that the energy produced in the electrical generator is efficiently transferred to the load, there is a well known and fundamental requirement that the impedance of the load should be significantly larger than the impedance of the generator. However, whilst this arrangement (Figure 2a) achieves maximum electrical efficiency (and prevents the generator from thermal destruction), it does not achieve maximum power transfer from source to load. Maximum power transfer occurs in the case where the load impedance is equal to the source impedance, as illustrated in Figure 2b. In the case of an AC energy source, the load should provide a conjugate match to the source. If the diagrams of Figure 2 were taken as a very basic representation of a conventional electromagnetic electrical generator supplying a load resistance, R_{Source} would represent the generator winding resistance and V_{Source} the EMF produced by time varying flux linkage with those windings.

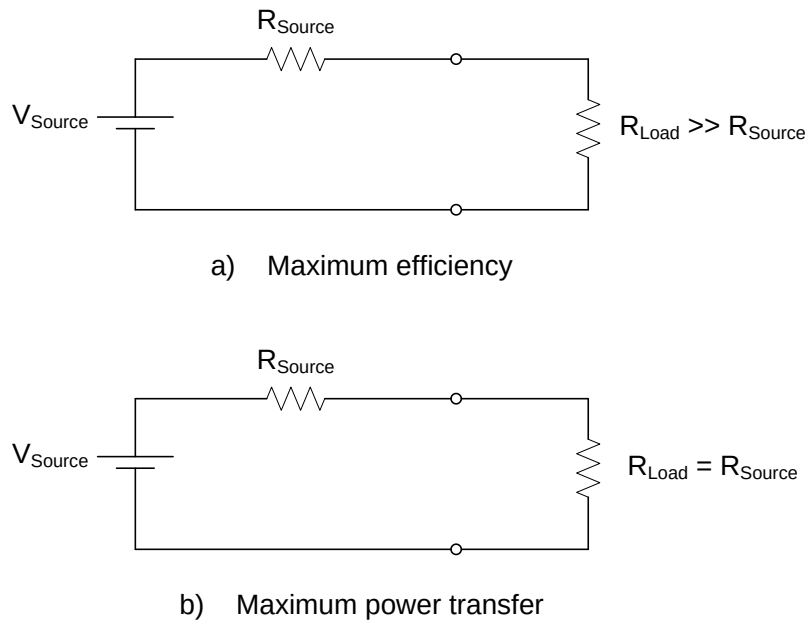


Figure 2 Maximum efficiency of energy transfer to load (a) and maximum power transfer to load (b)

In the case of energy harvesting systems, the fuel supply is effectively free and this leads to the desire to be able to transfer maximum power into the load, rather than to accomplish this at high efficiency. In addition, the quantities of power generated are low enough that an impedance match rarely has any thermal implications on the system.

In an energy harvesting generator, the definition of the impedance of the source to which the load should be matched is not generally as trivial as matching the load to a single electrical impedance. The source impedance will be dependent upon the type of energy harvester used and the conditions under which the harvester is operating. In some circumstances and harvester operating modes it may not be

optimal to match the impedance of the load to that of the source due to other constraints, however for energy harvesters studied in this chapter, there is always a clearly defined transducer load impedance which results in maximum power extraction from the transducer. It may therefore be more accurate to specify that the input impedance of the interface circuit to the transducer must be *controllable*, rather than always matched to the source, although in many cases the input impedance of the interface circuit will be set to match that of the source.

The details of source impedance modelling will be discussed in this chapter for each harvester type considered. The source impedance will always be shown as an electrical circuit which will often contain components which represent quantities other than pure electrical ones. As an example, vibration-driven harvesters, discussed in detail in Chapter 5, have a source model which takes into account the mechanical properties of the system such as the mass, the spring and the vibration characteristics as well as including the expected electrical resistance of the generator's windings or capacitance. All of these aspects must be included in the source model so that a suitable interface circuit can be designed, otherwise global system optimisation cannot be achieved [1].

Energy Storage

The vast majority of energy harvesting transducers will not be able to supply energy at a constant rate over long periods of time. Clearly a solar cell can only produce electrical energy when illuminated and a vibration-harvester only when it is subjected to an acceleration. However, many applications of energy harvesting technology may require a constant source of electrical energy to supply the load. If the average power consumption of the load is greater than the average power generated by the harvester, it is not possible to provide power continually to the load. However, if the average power generated is equal to or exceeds average consumption by the load, it is possible to run the load continually. However, in order to achieve this, the addition of a storage device, very likely electrical storage in the form of a battery or capacitor as discussed in Chapter 3, may be required.

Output Voltage Regulation

The many different types of energy harvesters produce power at different combinations of voltage and current. Photovoltaic cells and electromagnetic transduction kinetic harvesters tend to produce very low voltages (sometimes significantly less than 1 V) whilst electrostatic devices may produce their output power at over 100 V and potentially approaching 1 kV if operated optimally [2]. The output voltage from such devices must therefore be processed before being presented to the load electronics. In addition, if an energy storage element is included in the system, the voltage across that element may fluctuate depending on its state of charge. This effect may be negligible in the case of a storage battery, but may be significant if a capacitor is used as the storage component.

Overview

Often, the most difficult part of the harvester power electronics system to realise is the part which directly interfaces with the transducer, i.e. the part of the system that allows the generator to perform optimally through input impedance control. The implementation of this circuit is the part of the electronics that is most specific to each transducer technology used due to vastly differing voltage and current output combinations provided by the different transduction mechanisms.

The choice of storage, discussed in Chapter 3, and the output voltage regulation circuitry are generally common across all harvester systems with few characteristics being specific to the particular harvester type used. Therefore, the most harvester specific part of the electronics, the interface circuits with controllable input impedances, will now be discussed.

Interface Electronics for Kinetic Energy Harvesters

In order to determine an optimal electrical load for a motion driven harvester, a suitable source model must be developed, i.e. the impedance and output voltage characteristics of the source must be known. All aspects of the energy transfer (from vibration energy source through to the mass and spring and the transduction mechanism) must be taken into account in the source model. As the overall aim is to provide an optimal electrical load to the system, it is sensible to construct an electrical equivalent model of the generator which takes into account the mechanics of the system as electrical components. Two generic examples of such models are shown in Figure 3. A detailed explanation of the construction of these equivalent models is given in [3] and therefore only an overview will be given here.

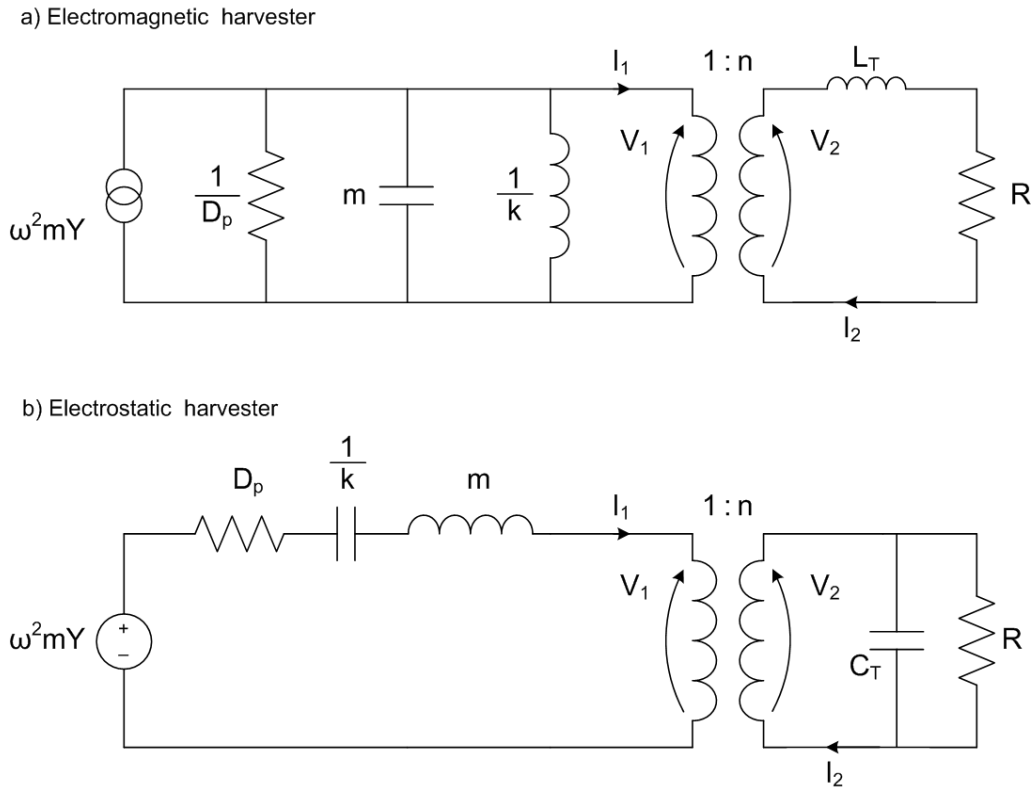


Figure 3 Equivalent circuit for motion driven harvester using electromagnetic force (a) and electrostatic force (b)

The circuits of Figure 3 show the equivalent circuit models for vibration-driven harvesters using electromagnetic damping (a) and electrostatic damping (b). The part of the circuit connected to the primary side of the transformer models the mechanical components. In a), the current source represents the input energy to the system (i.e. the mechanical vibration), the capacitor, m , represents the mass, the inductor, $1/k$, the spring and the resistor, $1/D_p$, the parasitic damping. In b), the voltage source represents the vibration source, the inductor represents the mass, the capacitor the spring and the resistor the parasitic damping. In both cases the transformer represents the coupling from the mechanical domain to the electrical domain through the transducer. In a), voltages across components on the left of the transformer represent velocity of those components and currents through them represent forces applied to them. The opposite is true for b). In both cases, the terminals on the secondary of the transformer represent the physical electrical connections of the transducer to which the interface circuit can be connected (in this case shown as a simple load resistor). The inductor, L_T represents the self-inductance of the coil in an electromagnetic device and C_T the terminal capacitance of either the piezoelectric material or the moving capacitor in the electrostatic device. It is important to note that the fundamental requirement for stored energy in these transducers places a limit on the maximum real power that can be transferred to a load resistor (in other words, energy stored in the inductance L_T or capacitance C_T). Whilst Figure 3a is a good model of an electromagnetic harvester and Figure 3b is a good model of a piezoelectric harvester, neither model is perfect for the electrostatic

moving capacitor transducer. This is because Figure 3 is a linear circuit and electrostatic transducers are inherently non-linear systems; their capacitance is non-constant.

The task, then, in the case of a motion-driven inertial generator, is to connect a value of load resistance (or much better, a power conditioning circuit feeding a storage element which together emulate a load resistance) which can absorb the maximum amount of energy from the energy source on the left of the transformer.

If we first assume that the storage elements C_T and L_T associated with the transducer have negligible effect, it is clear from Figure 3 that maximum power can be extracted from the source into the load (shown here as R) if the circuit is operated at a frequency where the inductor and capacitor resonate and if the load resistance equals the equivalent resistance of the parasitic damping when referred through the turns ratio. These models are therefore coherent with the analysis presented in Chapter 4, where it was concluded that maximum power is transferred to the load at resonance and when the electrical and parasitic damping are equal.

Therefore, in the case of our impedance match for a load to a motion driven micro-generator, the aim is often to produce a power converter which can feed energy into a storage element whilst maintaining an input impedance of resistance $1/D_p$. It should be noted that operating conditions exist where the optimal load resistance which should be presented by the interface circuit is not simply given by $1/D_p$. A different optimal resistance exists if the generator is operating off resonance and still a different expression can be found for the optimal resistance if the generator's proof mass becomes displacement limited, which may be the case if the parasitic damping can be made small. A comprehensive derivation of these different constraints is presented in [4]. However, whilst the optimal load resistance may change depending on the operating condition, in all these cases we conclude that there is an optimal impedance that should be presented by the power electronics interface circuit (Figure 4) to the electrical terminals of the micro-generator's transducer.

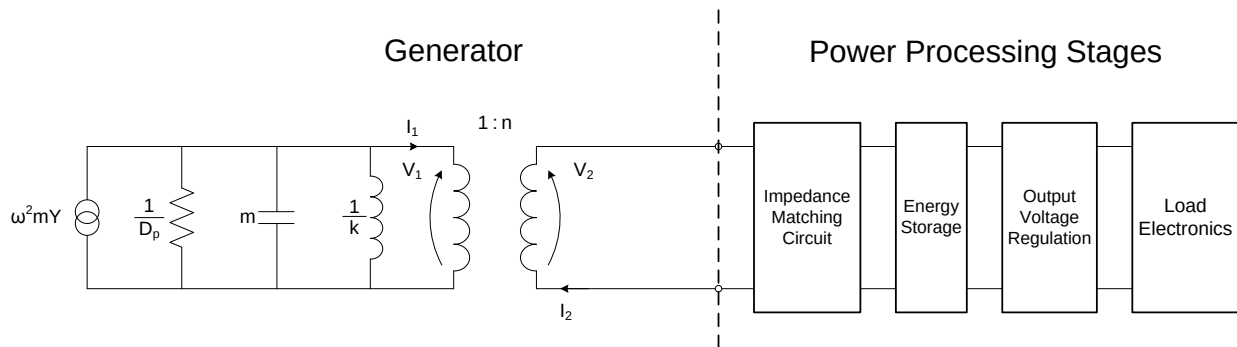


Figure 4 Connection of power electronics to electromagnetic generator model

We are now in a position to discuss specific implementations of electronics to interface with the three different transducer types for kinetic energy harvesters, i.e. electromagnetic transducers, electrostatic transducers and piezoelectric transducers.

Electromagnetic Harvesters

The general requirements for interfacing to an electromagnetic transducer on a vibration-driven micro-generator are:

- Rectification
- Voltage step-up capability
- Emulate a resistive load for the impedance match/impedance control

The simplest electrical interface for an electromagnetic harvester consists of a step-up transformer which feeds two Schottky diodes ($D1$ and $D2$) and a capacitor (C) which acts as a storage component, as shown in Figure 5 [5]. Due to the sinusoidal nature of the input vibrations, the output voltage from the electromagnetic harvester is AC. Using the transformer, the typically low transducer output voltage (tens or hundreds of mV) is up-converted through the use of the appropriate transformer turns ratio. Rectification of the stepped-up voltage is achieved by diode $D1$ which conducts during one half of the AC output voltage followed by $D2$ in the other half. This technique of using diodes to rectify the AC-voltages from vibration-based energy harvesters is quite common [6-8]. In the configuration shown in Figure 5, only one diode conducts during each half cycle of the input vibration when compared to a standard diode bridge thus minimising the effect of diode voltage drop, although this can still pose a problem. This configuration does not perform an impedance match between the electromagnetic harvester's source impedance and the interface electronics and therefore maximum power is not transferred from the harvester to the load. However, the simplicity of the arrangement in achieving rectification and voltage step-up is an advantage of this method.

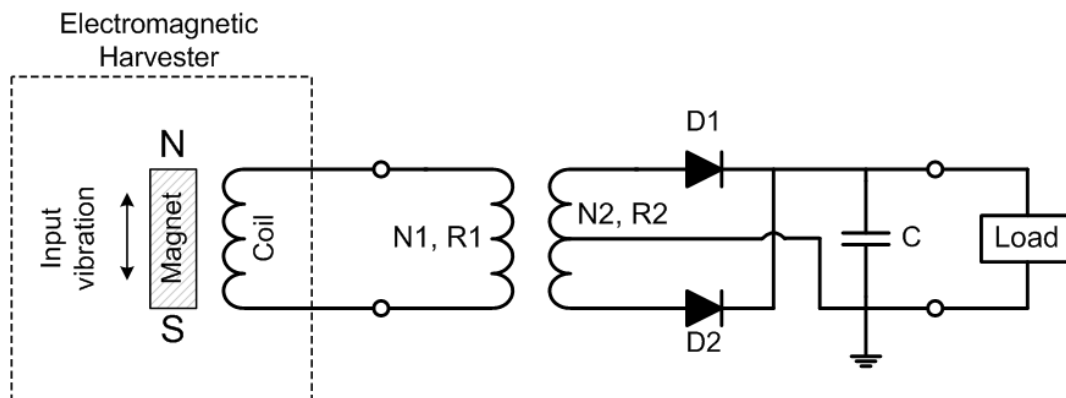


Figure 5 A simple electrical interface circuit which performs rectification and voltage step-up. (Redrawn from [5])

Alternatively, voltage multipliers such as the Villard multiplier (Figure 6) and the Dickson multiplier have been used to boost the voltage from the transducer. Cascading multiple stages of the Villard multiplier

will result in greater step-up ratios on the voltage from the transducer. One benefit of this approach over the previous arrangement is the ability to step up without using magnetic components, which favours integrated fabrication techniques. Again, such an approach fails to provide an impedance match.

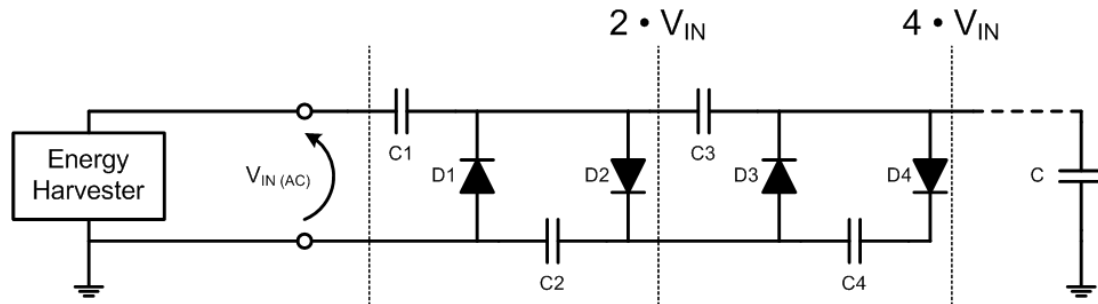


Figure 6 Using a Villard voltage multiplier for voltage up-conversion. (Redrawn from [9])

Mitcheson et al., proposed a dual-polarity boost converter that interfaces an electromagnetic generator in [1] as a potential solution to provide rectification, an impedance match and voltage step-up in one circuit, whilst minimising diode voltage drops. This converter provides low-voltage rectification of the positive and negative half cycles of the generated voltage: two boost converters are activated alternatively to rectify the AC voltage from the harvester's output. The dual-polarity nature of the converter removes the need for a diode bridge rectifier. Additionally, the circuit fulfils the step-up conversion requirements inherent on the output voltage of electromagnetic energy harvesters. Within the boost converter, the authors recommend the use of synchronously switched MOSFETs or Schottky diodes to reduce the effects of power losses in the converter.

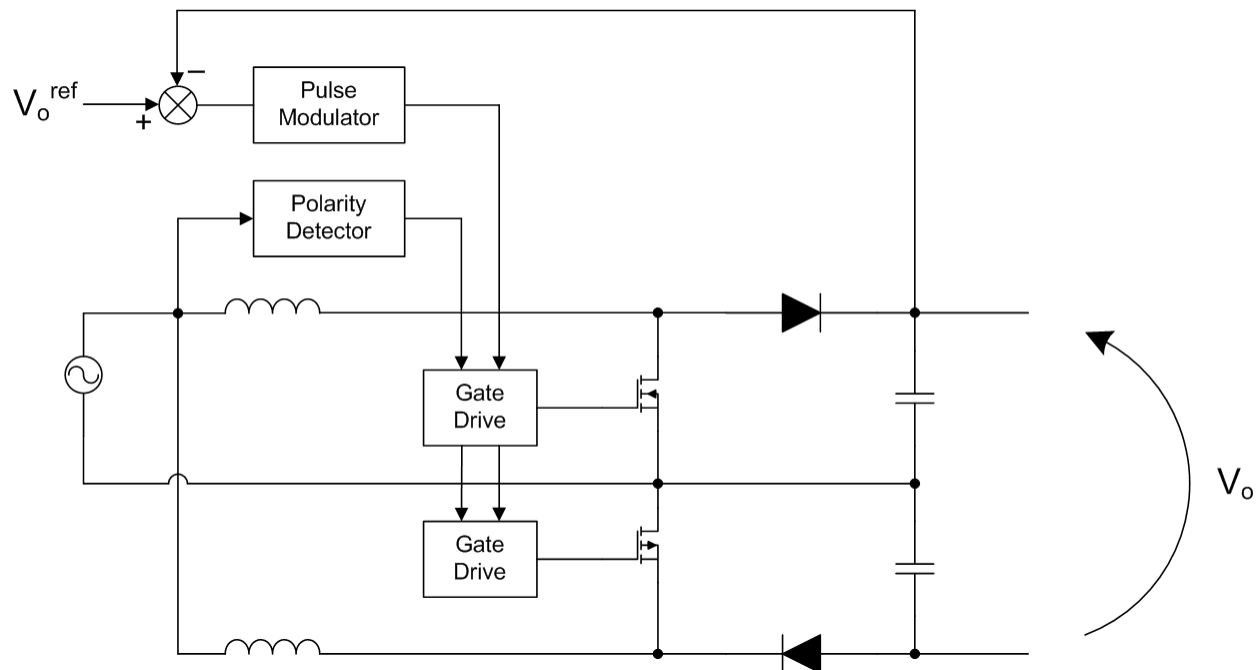


Figure 7 Dual polarity boost converter. (Redrawn from [1])

In [10], Maurath et al. reported on an adaptive impedance matching technique utilising switched capacitor arrays. The proposed circuit consumed less than $50 \mu\text{W}$ (simulated) and is geared towards self-powered applications for energy harvesters. Typically, output currents from microgenerators are quite low (less than 1 mA) which was why an on-chip capacitor-based impedance matching circuit was chosen to interface the generator. If the voltage across the switched capacitor array is half that of the generator's voltage, an impedance match exists between the generator's internal resistance and the load. This is an attractive impedance matching technique because it negates the need for current sensing within the power converter. The capacitors in the switched-array are charged to $(0.5V_{\text{gen}} + \Delta V_{\text{charge}})$ during a charging time period and then the switch toggles to the other state whereby the capacitors will then discharge to a storage capacitor which feeds a boost converter. At the end of the discharge cycle, the voltage across the capacitor array will decrease to $(0.5V_{\text{gen}} - \Delta V_{\text{discharge}})$. The switching frequency for these capacitor arrays depends on how small the ΔV 's are required to be and hence is closely linked to the efficiency of the circuit. The control of the circuit is not described in [10] in detail but it is likely that some open circuit measurement of the transducer open circuit voltage would need to be made during operation as the operating conditions change.

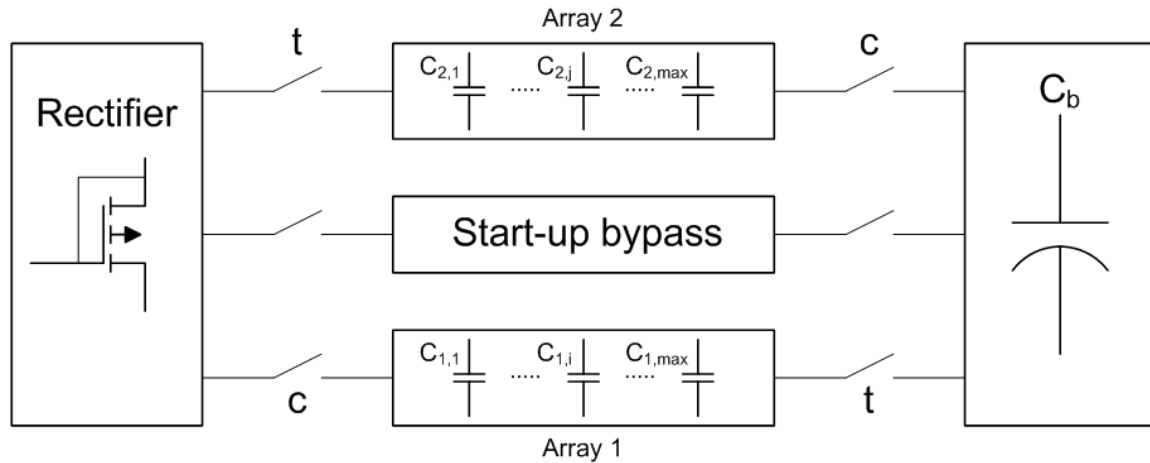


Figure 8 Adaptive impedance matching technique using switched capacitor arrays. (Redrawn from [10])

Example Complete Power Electronics System for Continually Rotating Energy Harvester

Many examples have been presented in the literature and, indeed, earlier in this book, about vibration powered harvesters. High performance power electronics with all the functionality of optimal damping control (the impedance match), energy storage and output voltage regulation, have yet to be demonstrated for such systems (mainly because of the difficulty of achieving these functions with such low power generation capability and the need that these functions must be powered from the energy generated (although simulations of some or all of these aspects has been demonstrated). However, all of these functions have already been practically demonstrated for a different type of energy harvesting device: the rotational harvester based on gravitational torque. This harvester is implemented with an electromagnetic transducer and therefore many of the features required for the vibration case are shared with the rotational case. Here, then, we will look in some detail about the design and realisation of the complete power electronic system, described in Figure 1, for this kind of harvester.

The operation of the gravitational torque harvester is as follows: the rotor of a conventional electrical generator is connected to a rotational host source from which energy is being harvested. As the rotor spins, the stator is held in position by the force of gravity acting on an offset counterweight on the stator, as shown in Figure 9(a). As current is drawn from the generator, the torque between the rotor and stator is counteracted by the gravitational torque on the offset mass and power is generated. Another possibility for configuring the generator is shown in Figure 9(b) where the stator of the generator is connected to the host and the offset mass is attached to the rotor of the generator. Detailed operation of these devices is described in thoroughly in [11] and [12].

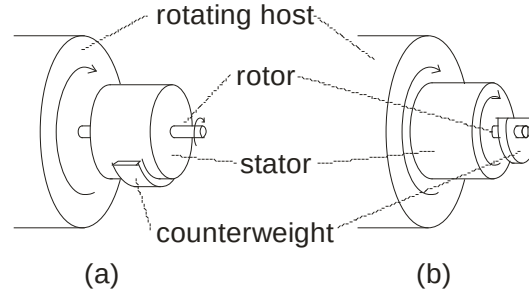


Figure 9 Two possible configurations of a rotational harvester constructed from a DC motor: (a) the offset mass is attached to the stator and the rotation is coupled to the rotor or (b) the offset mass is attached to the rotor with the rotation coupled to the stator. (Redrawn from [11])

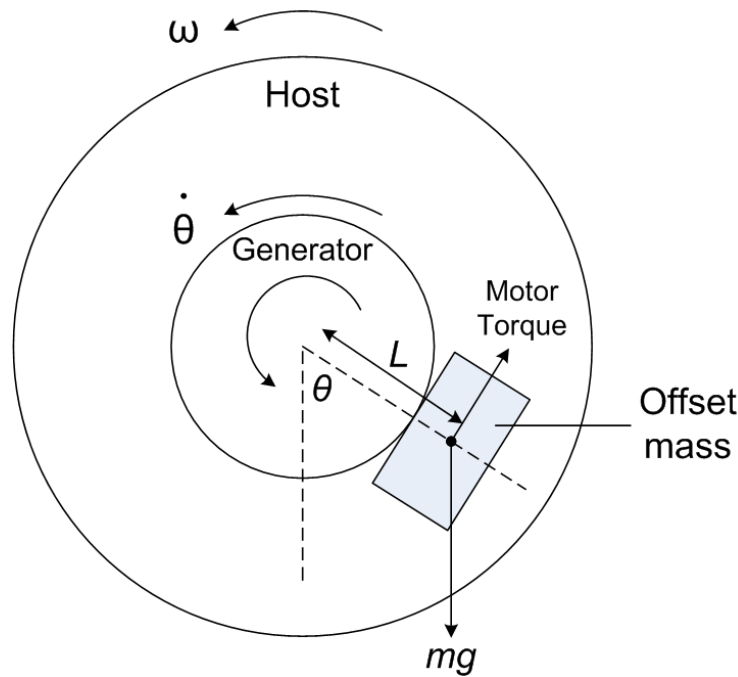


Figure 10 End view of rotational torque harvester. (Redrawn from [11])

As current is drawn from the rotational harvester, a torque causes the proof mass to rotate such that the torque from gravity, $T_g = mgL \sin(\theta)$, counteracts the motor torque, as shown in Figure 10. For a given rotation speed ω of the host, the limit on the electrical power that can be generated is given by $T_g \omega$, assuming that the mass is held at 90° to the vertical. If the angle of the offset mass exceeds 90° , the rotor and stator of the generator will start to synchronise and power generation will be substantially reduced. From this basic argument it seems that a current should be drawn from the generator such that the angle of the mass is held at 90° . However, when we consider the amount of that power that can be dissipated into a load, or pushed into a storage element, (in other words the useful electrical power) we must consider the electrical equivalent circuit of the generator and load as shown in Figure 11, whilst also considering the constraints of the mechanical system. It is clear that, for a given rotation speed and therefore value of open circuit generator voltage E_g , maximum power will be transferred to a load which

is matched to the impedance of the armature, R_{ARM} . There are therefore two operating modes for this system to ensure maximum power is generated:

- At low rotation speeds, the impedance of the load should be equal to the generator armature resistance. In this mode the load resistance is constant.
- As the rotation speed of the host increases under matched conditions, eventually the offset mass will reach 90° . At this point the load impedance should be increased to prevent the mass flipping and the synchronisation of the generator's rotor and stator. Therefore in this operating mode the generator current should be held constant.

The input impedance of the interface circuit must therefore be controllable to ensure maximum energy can be harvested under all operating conditions.

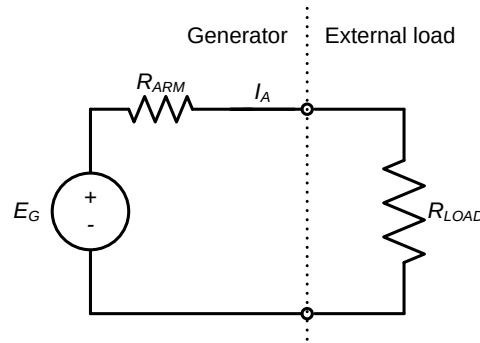


Figure 11 Simple DC model of generator

As explained previously, we do not want to simply dissipate power in a load resistor, but to supply power to charge a storage element and to power useful loads. Consequently, a power electronic system must be designed which is able to charge a storage element and to present either a constant impedance (at low rotation speeds) or constant current sink interface (at high rotation speeds) to the generator.

The overall topology for the power electronics is therefore as shown in Figure 12.

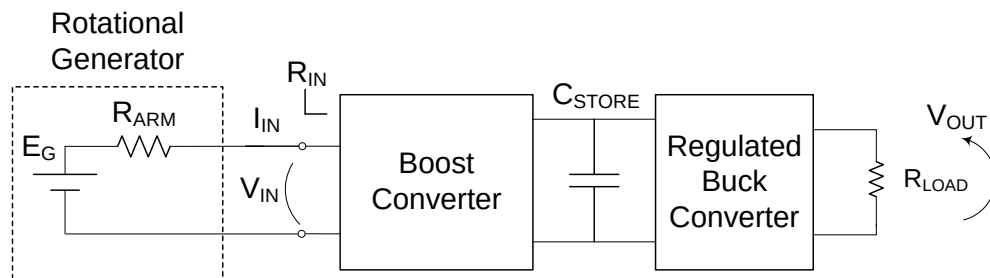


Figure 12 Power processing topology for rotational harvester. (Redrawn from [11])

A boost converter was chosen as the interface to the generator because it is able to provide smooth input currents (and thus emulate a resistive input impedance) and step up the relatively low voltages from the generator to push energy into a capacitor, which acts as an energy store able to supply current to the load and smooth out the intermittency of the generation of harvested energy. The voltage on the capacitor will rise if the rate of generation exceeds consumption by the load and vice versa. The final stage is a step-down converter which regulates the voltage for use by the load circuit.

A typical rotational harvester may be able to generate around 100 mW, depending on its size and the rotation speed of the host. At these power levels, wide input voltage encapsulated switch mode converters with output voltage regulation are available off the shelf at low cost and with high efficiency. Therefore, the final stage of the system shown in Figure 12 is readily available for this system. The storage element can simply comprise super capacitors. However, a boost converter (Figure 13) with the right characteristics (i.e. input impedance control or input current control) is not readily available and must be designed. The design, construction and test of this converter will now be discussed.

Boost Converter Design

The design of power converters which process power in the range of a few Watts would normally involve a relatively standard procedure of choosing a switching frequency and inductor combination that would give an adequately low current ripple, choosing a large enough output capacitor to reduce output voltage ripple and then a diode and MOSFET with suitable voltage and current ratings and switching speeds [13]. However, in the design of a power converter for processing small amounts of power, the overhead of the control circuitry must be taken into account. In such a converter it is also desirable to reduce component count for simplification and in an attempt to reduce power consumption, and therefore use of components such as a separate gate drive and active filtering of feedback signals should be minimised. In addition, at these low power levels, the energy required to charge the gate capacitance of the MOSFET should be taken into account, as it may constitute a significant proportion of the energy loss in the converter. These additional issues make the optimisation of the converter more complicated. The design steps presented in this section assumes the reader has a basic knowledge of operation of switch mode power converters and does not cover the mathematical analysis of basic boost converter operation. A detailed introduction and analysis of the switch mode power converters described in this chapter can be referred to in [13]. Here, we focus on the exact component choices in order to maximise the efficiency of the converter for an energy harvesting system and to allow the converter to work at low input voltage.

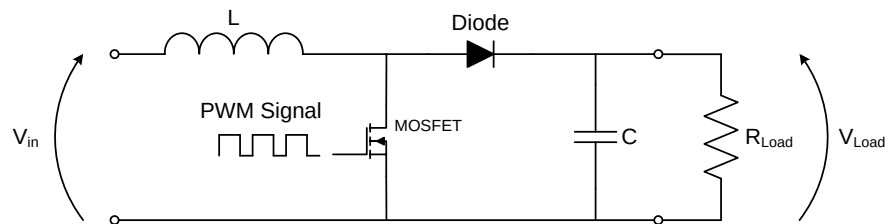


Figure 13 Boost converter used as interface circuit to transducer

The approach taken for this design was to optimise the boost converter around what we considered to be a likely operating point for the system, shown in Table 1.

Table 1 Operating point for optimisation

Generator output impedance	9.1 Ω
Generated EMF from transducer	4.4 V
Capacitance of Energy Storage	2 mF
Storage capacitor nominal voltage	15 V

The individual power losses in the circuit, whose sum should be minimised, are given in Table 2.

Table 2 Loss mechanisms in boost converter

Inductor conduction loss
Diode conduction loss
Diode reverse recovery loss
MOSFET conduction loss
MOSFET switching loss
MOSFET gate charge energy loss

There are several free parameters that can be chosen in order to attempt to minimise energy loss in the circuit. These are listed in Table 3.

Table 3 Design parameters

PWM Switching frequency
Inductor current rating

Inductor inductance
MOSFET voltage and current rating
Diode current and voltage rating

Unfortunately, changing one parameter to reduce one of the losses can cause an increase in other losses. For example, increasing the diode current rating in order to reduce diode conduction loss will almost certainly increase diode reverse recovery losses and therefore a complete system optimisation (accounting for all the parameters at the same time) must be performed.

Expressions for the power losses shown in Table 2 were derived in terms of the operating point of the converter and the design parameters of Table 3. As an example, the derivation of formulae for the transistor's conduction loss, switching loss and gate charge energy loss now be described.

Conduction Losses

Conduction losses are dependent on the drain-source resistance R_{DS} of the transistor and are proportional to the square of the Boost converter's input current multiplied by the duty cycle. The two free design parameters for the MOSFET are the current rating and voltage rating. The maximum voltage blocking capability required by the MOSFET in this case was 40 V as this was the breakdown voltage of the storage capacitance. As, under a given operating current, conduction loss in a MOSFET is approximately proportional to the square root of the maximum blocking voltage [14], it makes sense to use a MOSFET with the rating required by the application without over-rating the device's voltage. This means that the best device for the application is a 40 V MOSFET whose current rating must be determined. Initially, R_{DS} values were gathered for a range of 40 V MOSFETS as a function of their rated operating current, as shown in Figure 14.

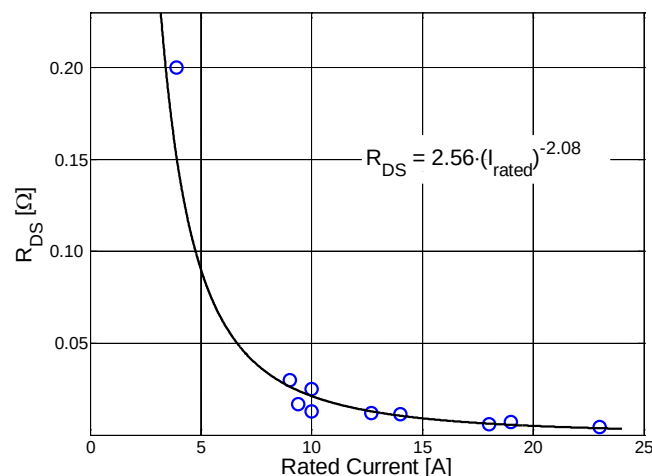


Figure 14 Relationship between the drain-source resistance, R_{DS} and the transistor rated current.

By applying a curve-fit to the points, a relationship between R_{DS} and rated current was obtained as:

Equation 1
$$R_{DS} = 2.56 \cdot (I_{rated})^{-2.08}$$

The conduction loss can then be expressed as:

Equation 2
$$P_{cond} = \delta \cdot I_{in}^2 [2.56 \cdot (I_{rated})^{-2.08}]$$

Switching Losses

Switching losses arise from the fact that the MOSFET takes time to switch on or off, fundamentally because it takes time to push charge on and off its gate. The time taken to switch between these two states depends on the stray capacitances at the gate-source and gate-drain junctions, C_{GS} and C_{GD} respectively, and the current drive capability of the gate drive circuitry. Values of these capacitances are always provided in the datasheets but as they are voltage dependent it is better to perform calculations based on gate charge (Q_{GS} and Q_{GD}) instead of capacitance.

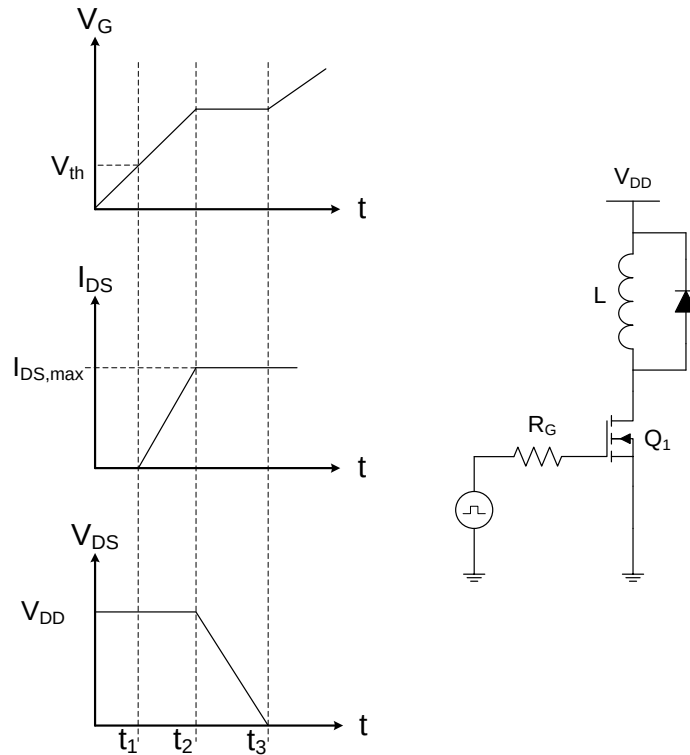


Figure 15 Typical voltage and current waveforms as the transistor turns on to switch an inductive load.

Figure 15, shows the typical waveforms of a MOSFET switching an inductive load, as is the case in this circuit. This waveform is described in some detail in [15]. When the gate drive source of the FET is initially set high, V_G begins to increase until it reaches the threshold voltage V_{th} of the FET at time t_1 . At this point, the drain current I_{DS} starts to increase. C_{GS} continues to charge until the drain current is equal to the inductor current at t_2 . At time t_2 , V_G and I_{DS} remains constant as the Miller capacitance, C_{GD} , is charged. At t_3 , the FET is fully switched on and the voltage drop across the drain-source region is almost negligible. V_{GS} then stabilises at its final value.

Power loss due to switching occurs in the period between t_1 and t_3 , where there is both a non-negligible current through the MOSFET and non-negligible voltage across it. The instantaneous power loss is shown in Figure 16.

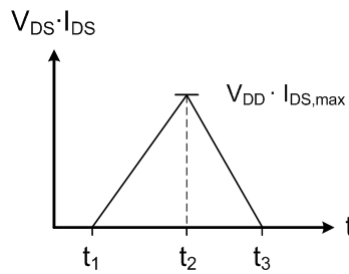


Figure 16 Switching power loss waveform

Equation 3
$$P_{SW} = \frac{1}{2} (V_{DD} \cdot I_{DS,max}) \cdot (t_3 - t_1) \cdot f_{sw}$$

V_{DD} and $I_{DS,max}$ are known operating conditions for the converter, and so in order to calculate switching loss, only t_1 and t_3 must be found. In our example, the gate drive for the MOSFET is an output pin on a PIC18F1320 microcontroller [16]. As discussed above, the time taken for switching is the time taken to charge C_{GS} and C_{DG} . The current to do this is supplied by the PIC and the output pin on the 18F-series is capable of driving 25 mA.

Therefore, the switching times can be estimated from:

Equation 4
$$t_1 = \frac{Q_{GS}}{I_{PIC}} = \frac{Q_{GS}}{25mA}$$

Equation 5
$$t_3 = \frac{Q_{GD}}{I_{PIC}} + t_1$$

Values of Q_{GD} and Q_{GS} can be estimated from the plots of gate-source voltage against total gate charge given in the datasheets (Figure 17). It is possible to correlate the individual gate charges to the time instances t_1 to t_3 . For example, t_1 is the time required to raise the gate voltage to the threshold voltage, t_2 is the time at which C_{GS} is sufficiently charged to support the drain current set by the inductor and the interval from t_2 to t_3 is the time taken to charge the Miller capacitance, C_{GD} .

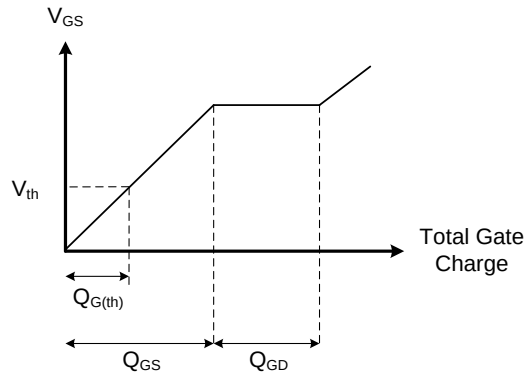


Figure 17 The charging of C_{GS} and C_{GD} depends on the applied V_{GS}

By inspecting the plots of gate voltage against total gate charge, values of $Q_{G(th)}$, Q_{GS} and Q_{GD} were estimated for each transistor, along with their respective rated currents (Figure 18).

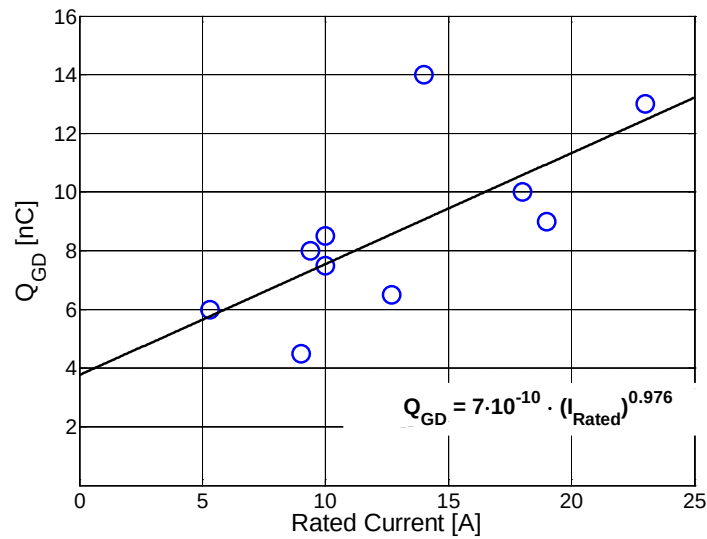


Figure 18 Estimated Q_{GD} values as a function of rated current.

The expression relating Q_{GD} and rated current was found to be:

Equation 6
$$Q_{GD} = 7 \times 10^{-10} \cdot (I_{Rated})^{0.976}$$

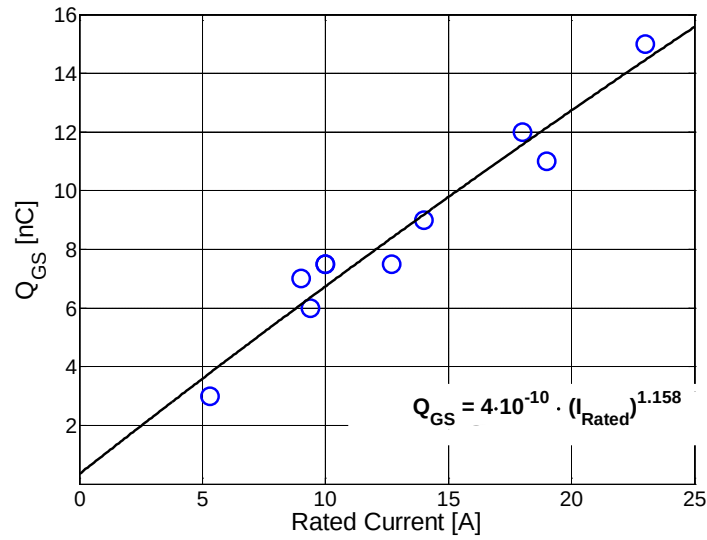


Figure 19 Estimated Q_{GS} values as a function of rated current.

The relationship between Q_{GS} and rated current is:

Equation 7
$$Q_{GS} = 4 \times 10^{-10} \cdot (I_{Rated})^{1.158}$$

Finally, the analytical expression for switching power loss is given by:

Equation 8
$$P_{SW} = \frac{1}{2} (V_{DD} \cdot I_{DS,max}) \cdot [4 \times 10^{-10} \cdot (I_{Rated})^{1.158}] \cdot f_{SW}$$

Gate Charge Losses

The stray capacitances C_{GD} and C_{GS} are repeatedly charged and discharged during the turn-on and turn-off transients when switching transistor Q_1 (Figure 20). This causes energy loss as none of the energy placed on these capacitors is ever recovered.

Figure 20a shows a switching circuit with gate drive from which the flow of charge through these stray capacitances can be analysed and thus energy losses calculated. Transistors T_1 (PMOS) and T_2 (NMOS) were assumed to be ideal switches in a gate drive, V_G is the gate drive power supply voltage, and R_G is the output resistance of the gate drive.

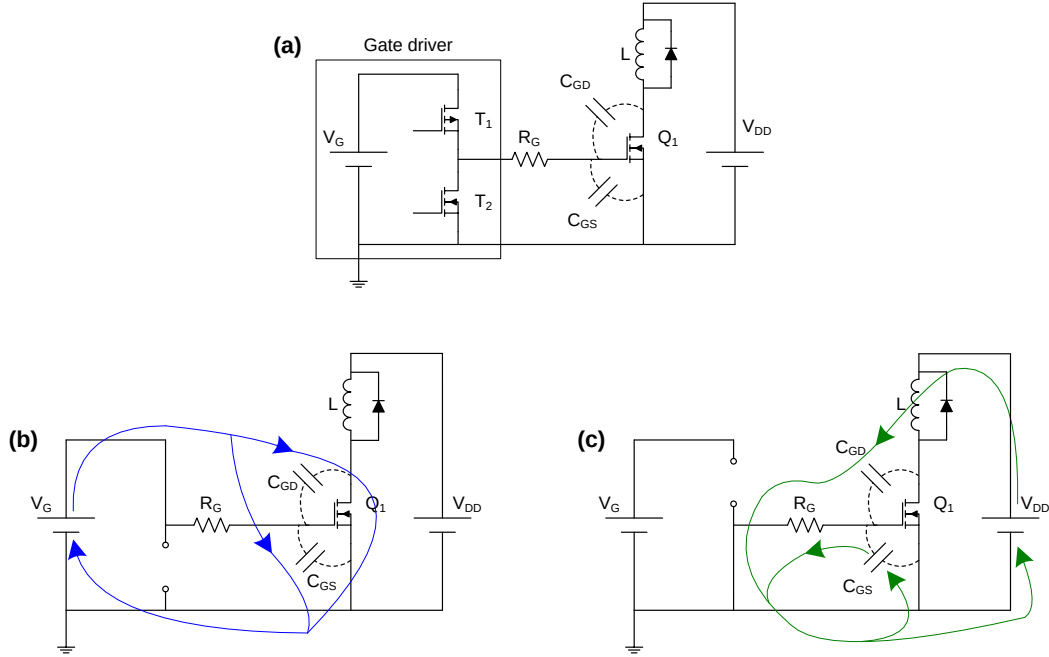


Figure 20 Current flow through C_{GD} and C_{GS} during the turn-on (b) and turn-off (c) transients.

The gate charge energy loss occurs in two instances: when the transistor is being switched on and when it is being switched off. Power loss due to the gate capacitance occurs when charge is taken from V_G or V_{DD} to bias C_{GS} or C_{GD} . Consider the turn-on scenario in Figure 20b where T_1 becomes a short circuit and T_2 is open circuited. Energy is transferred from the gate drive supply, V_G to C_{GD} and C_{GS} as indicated by the flow of currents as shown by the arrows. Capacitor C_{GS} is charged by the gate driver from zero volts to V_G . Also, when Q_1 turns on, its drain voltage must fall from V_{DD} to ground. To achieve this, C_{GD} would have had to accumulate charge from V_G and this amounts to an energy of $(Q_{GD} \cdot V_{GS})$. The amount of power lost in the stray capacitances is therefore given by:

$$\text{Equation 9} \quad P_{Gate(ON)} = f_{SW} \times (Q_{GS} V_{GS} + Q_{GD} V_{GS})$$

As the transistor is switched off, both capacitances will discharge according to the path shown by the arrows shown in Figure 20c. Here, T_1 is open circuited and T_2 is shorted to ground. The current from C_{GS} will flow directly to ground (and thus no further energy is taken from a voltage source) whereas V_D supplies the energy to bias C_{GD} in a direction opposite to that in Figure 20b. Therefore, work is done to raise the voltage on the drain from zero to V_D . This gives a turn off power loss of:

$$\text{Equation 10} \quad P_{Gate(OFF)} = f_{SW} \times (Q_{GS} V_{GS})$$

Consequently, the total power loss due to the gate charges is the sum of Equation 9 and Equation 10:

$$\text{Equation 11} \quad P_{Gate} = f_{SW} \times [Q_{GS} V_{GS} + Q_{GD} (V_{GS} + V_{DS})]$$

Total Transistor Power Loss

Adding all the power loss expressions together gives the total power loss in the transistor as a function of the device's rated current and switching frequency:

Equation 12
$$P_{FET} = P_{Cond} + P_{SW} + P_{Gate}$$

Optimisation in Matlab

By applying the same approach to finding expressions for losses in the diode and the inductor, an analytic expression for the complete power loss in the circuit was found as a function of the variables in Table 3. Then, the Matlab function *fmincon* was used to find the minimum value of the total power loss. The results are shown in Table 4.

Variable	Value
FET Rated Current	4.52 A
Diode Rated Current	0.56 A
Switching Frequency	36.2 kHz
Inductance	0.8 mH

Table 4 Results from the minimization process

These results were validated by sweeping each variable to ensure minimum power loss (and thus maximum useful output power) resulted from these specific values. The graphs in Figure 21 confirm that the minimization process was accurate in that it had found a minimum power loss for the system. As a result, the boost converter interface circuit was built using components which were chosen based on the values given in Table 4.

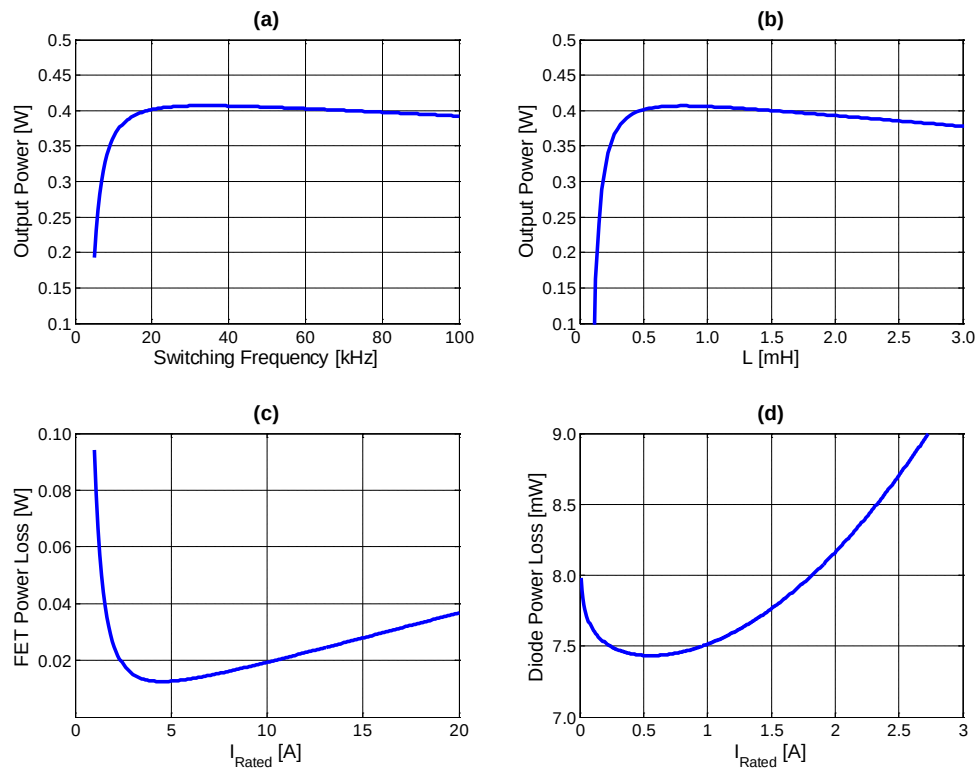


Figure 21 Validation of the optimisation process

Performance of the Boost Converter

The efficiency of the prototype boost converter was found to have peak values of approximately 96% for duty cycle values less than 0.80. The pulse-width-modulated (PWM) signal was provided by an external signal generator and was setup such that the PWM frequency (36.2 kHz) and peak-to-peak (3.3 V) values were the same as the microcontroller would provide. In addition to that, a 500 Ω load resistor was connected to the output. At higher duty cycle values, more current flows in the boost converter causing an increase in I^2R losses which degrades the efficiency of the converter as shown in Figure 22.

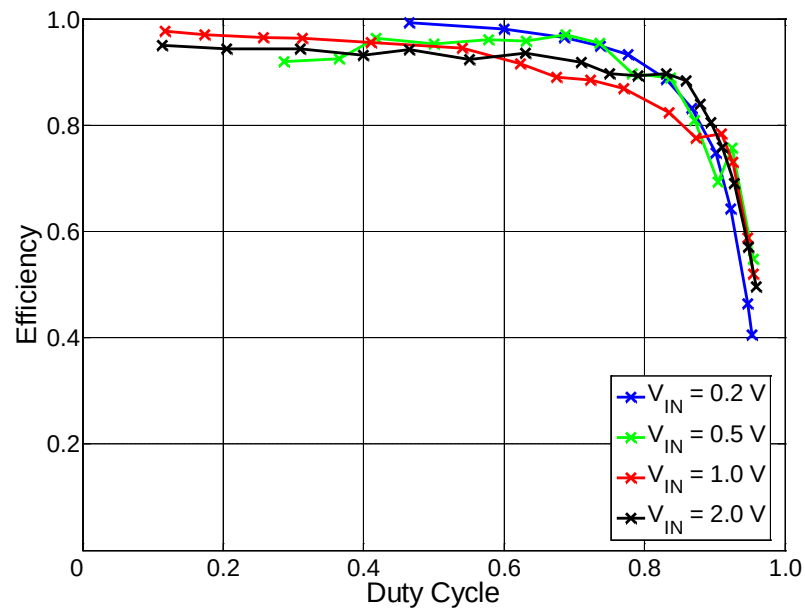


Figure 22 Efficiency of the boost converter at various duty cycle values.

Characterisation of the boost converter DC transfer characteristic was performed by applying various input voltages (0.2 V, 0.5 V, 1.0 V and 2.0 V) with a 500 Ω load resistor connected at the output. A maximum voltage gain of 11.1 was achieved at a duty cycle of 0.95 for an input voltage of 0.5 V. The experimental results follow the ideal voltage gain (grey crosses) very closely as shown in Figure 23.

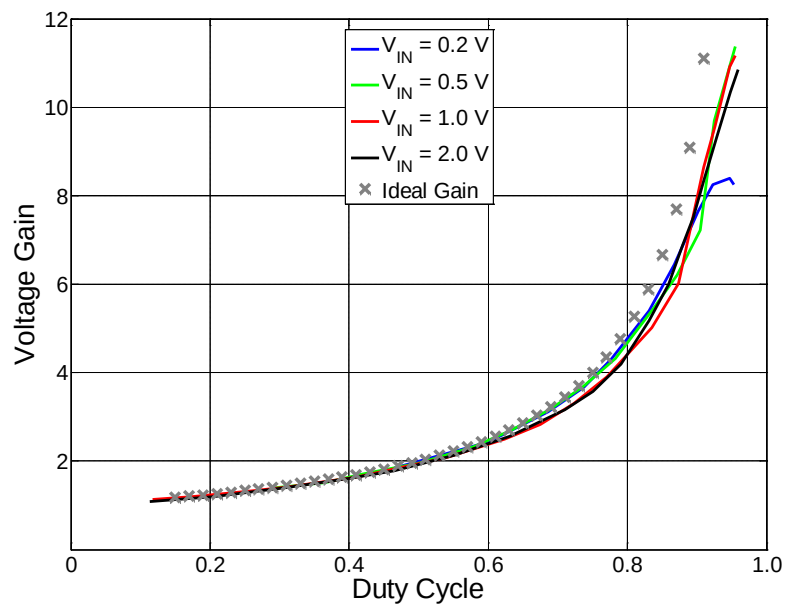


Figure 23 Voltage gain characteristics of the boost converter, at different input voltages.

Input Impedance Control

As previously discussed, optimal power transfer from the generator to a load requires that the load resistance, R_{LOAD} , matches the generator's armature resistance, R_{ARM} , when the generator's offset mass is held at less than 90° to the vertical and that the current be controlled to a maximum value when the offset mass reaches 90° (Figure 10). The input impedance, R_{IN} , of a boost converter can be altered to be less than its load impedance R_{LOAD} by varying its duty cycle, δ . It was assumed that the value of R_{ARM} would be relatively small compared to the input impedance of a device that would potentially be powered by this generator.

Equation 13
$$R_{IN} = R_{LOAD} \cdot (1 - \delta^2)$$

The flow chart in Figure 24 demonstrates a conceptual implementation of a boost converter to perform this impedance match.

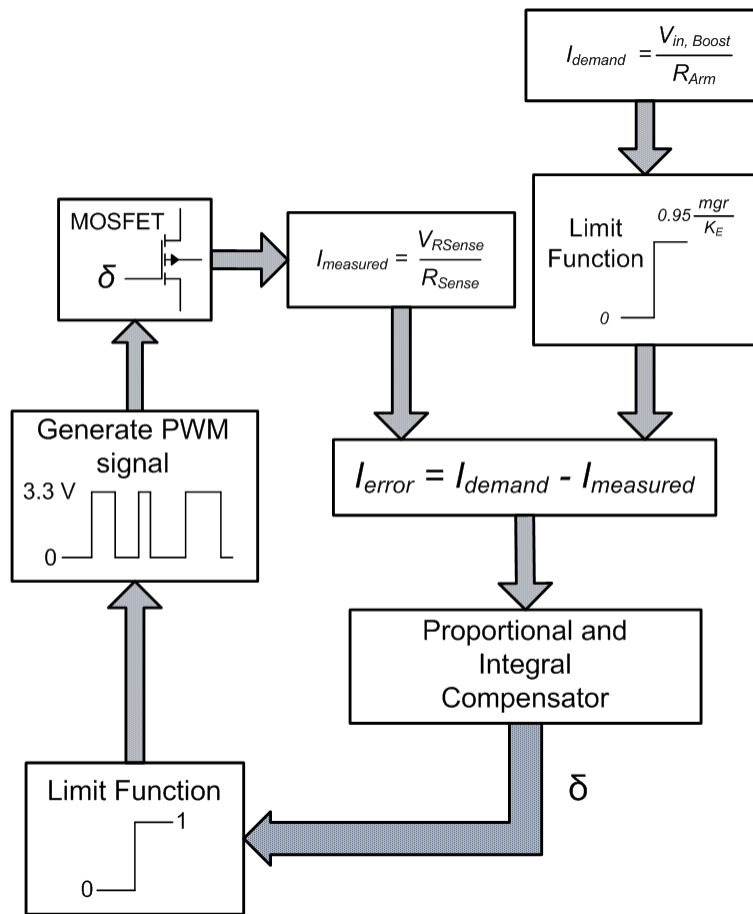


Figure 24 Flow chart of the boost converter input impedance matching procedure.

The boost converter inductor current can be measured using a sense resistor R_{sense} and a current sense amplifier. Since the inductor current is the armature current from the generator, the on-line optimisation procedure will match this inductor current to a demand value. This current demand value is obtained from the boost converter's input voltage, divided by the armature resistance, which is measured offline. This gives an indication of how much inductor current should be flowing in the circuit in order to present a near perfect impedance match between the generator's armature resistance and the load resistance that the generator sees. The error between the two currents is sent to a proportional and integral (PI) compensator which calculates the duty cycle required to match the measured current as close as possible to the demand current.

Circuit Implementation

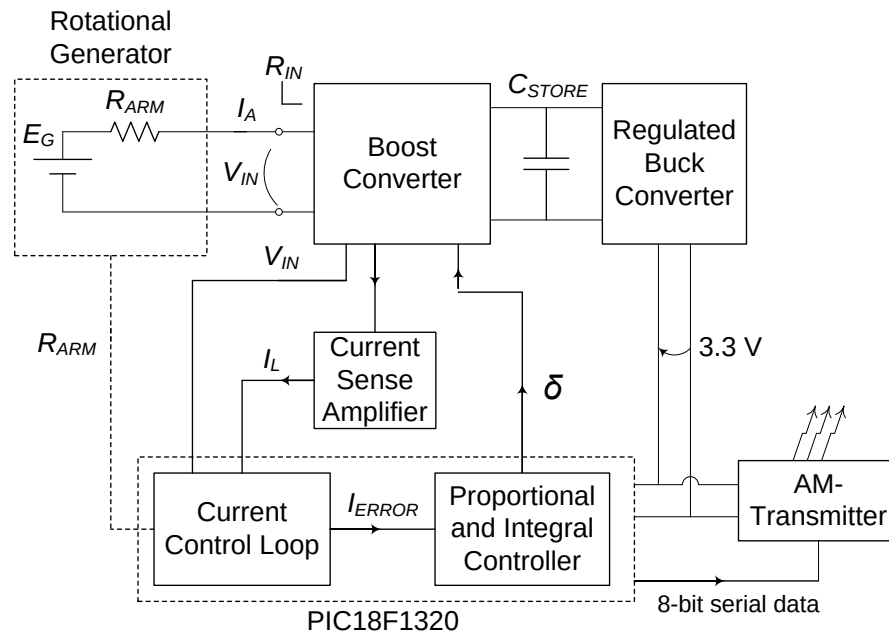


Figure 25 Schematic of the power processing and control circuitry. (Redrawn from [12])

Figure 25 shows a block diagram of the power processing and control circuitry that implements the impedance match. A storage capacitor C_{STORE} was placed between the boost converter and an off-the-shelf RECOM regulated buck converter allowing accumulation of energy and output voltage regulation respectively. C_{STORE} consists of three series-connected 6 mF supercapacitors rated at 15 V, from AVX. The buck converter has a wide input range (4.75 V – 34 V) and a regulated output (3.3 V) so that an external device can be powered at a fixed voltage of 3.3 V. The microcontroller samples the boost converter's input voltage, inductor current and the voltage across C_{STORE} whilst generating the required duty cycle to perform an impedance match. An AM-Transmitter from RF Solutions, operating at a

bandwidth of 433 MHz, was used to transmit the voltage levels of the storage capacitor to a PC, thus implementing a self-powered wireless sensor node.

Impedance Matching Results

The control loop outlined in Figure 24 was verified using a power supply to mimic the input voltage and current to the boost converter whilst a series connected resistance ($9.1\ \Omega$) was used to simulate the armature resistance of the generator. Two load resistance values ($50\ \Omega$ and $100\ \Omega$) were connected in parallel with the storage capacitor while the boost converter's input voltage was varied from $0.3\ \text{V}$ to $2.0\ \text{V}$. The input current changes proportionally with the variations in input voltages. The gradient of the graph in Figure 26 shows that the input impedance was held at $9.1\ \Omega$, for both load resistances.

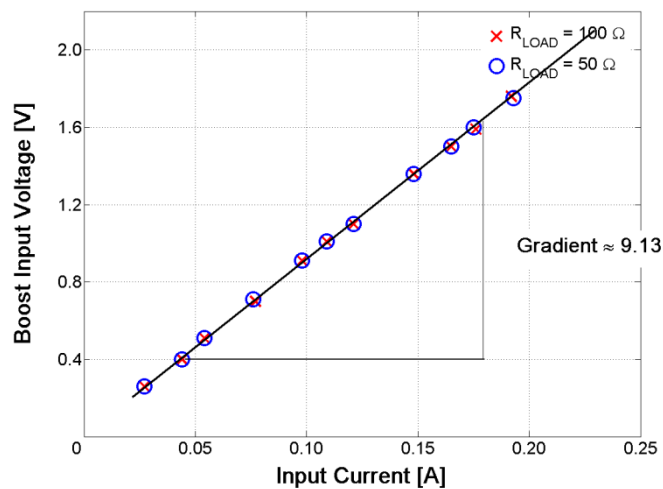


Figure 26 Impedance matching performance of the current control loop. (Redrawn from [11])

Figure 27 shows the results obtained when two different load resistances ($50\ \Omega$ and $100\ \Omega$) were connected to the output of the boost converter. The graphs illustrate changes in duty cycle and correspondingly the storage capacitor voltage while the input impedance of the boost converter was continuously matched to the target armature resistance of $9.1\ \Omega$.

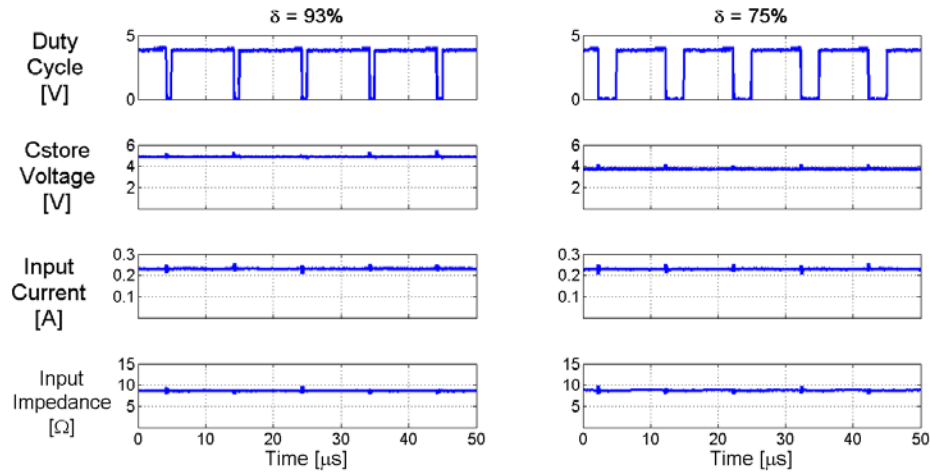


Figure 27 Variations in duty cycle under different load resistances to achieve an input impedance of 9.1Ω . (Redrawn from [11])

In Figure 28, a varying input voltage was applied to simulate a condition where the rotation speed of the generator changes. It was observed that the input current changes proportionally to the input voltage in order to maintain a fixed input impedance of 9.1Ω . When the generator's speed increases, more power is generated than is consumed by the load, leading to an increase in the voltage across the storage capacitor. When the contrary happens, the storage capacitor will discharge to maintain the operation of the impedance matching circuit. For the whole time, the output voltage from the Buck regulator stays at the predetermined value of 3.3 V and as importantly, the input impedance stays matched to R_{ARM} – an essential requirement for harvesting energy optimally from a rotational source under practical situations.

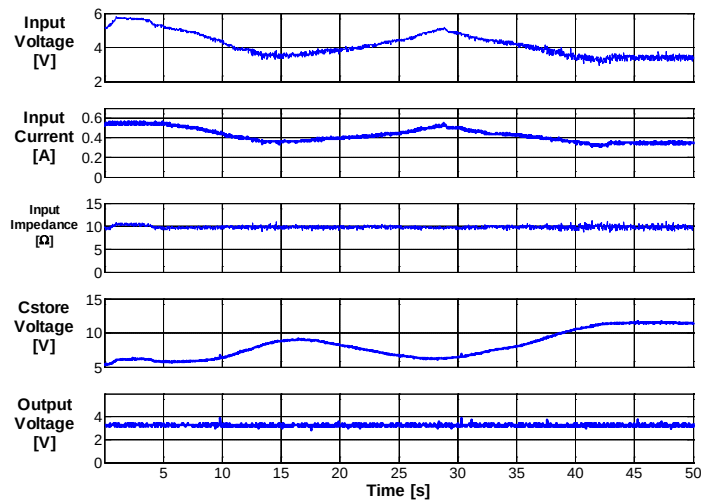


Figure 28 Performance of the impedance matching circuit for a varying input voltage and fixed load. The input impedance remains matched to R_{ARM} , 9.1Ω . (Redrawn from [11])

Conclusions for Power Electronics System for Continually Rotating Harvester

A power electronics system for an energy harvester which includes a transducer interface circuit, energy storage and output voltage regulation has been developed and demonstrated. The main difficulty in the design is that the circuit must be efficient, operate over wide voltage ranges and the control circuit must consume very little energy so that the system is capable of being self-sustaining from the harvested energy whilst still being able to supply power to a load. An end-to-end system optimisation was described for a boost converter interface circuit and this minimised the losses in the converter, resulting in an efficiency of 96%. The overall aim was to provide an impedance match to the generator's armature resistance and at the same time supply a regulated output voltage from which a load can be powered from whilst storing energy to allow the system to maintain operation when the energy harvesting source is intermittent. Therefore, all three functions required for an energy harvesting system, i.e. transducer interfacing for maximum power extraction, energy storage and output voltage regulation have been demonstrated in the above example.

Piezoelectric Harvesters

The typical electrical equivalent circuit of a vibration-driven piezoelectric harvester is shown in Figure 3b. When previously considering the design of interface circuits for electromagnetic devices shown in Figure 3a, we noted that in order to maximise power extraction from the transducer we should set the interface circuit to have an input impedance of $\frac{1}{D_p}$, assuming that the generator was operating at resonance and that no other constraints (such as displacement limit of the mass) were in operation. This argument is valid as long as the inductance of the transducer is negligible and this is frequently the case for the electromagnetic harvester (although not always). However, for piezoelectric transducers, the shunt capacitance can never be neglected because of the low coupling coefficient of the piezoelectric material.

A poor coupling between the mechanical and electrical domains of the piezoelectric material means that the transformer component in Figure 3b is a step up transformer with a high turns ratio. This means that very little voltage is developed across the primary side of the transducer. Therefore, at resonance, the mechanical motion of the transducer (i.e. its maximum displacement) is set almost entirely by the mechanical parasitic damping on the primary side of the transformer rather than the electrical loading. As a consequence, the piezoelectric current generated is almost independent of the electric loading on the generator and the equivalent circuit can be replaced with a much simpler model as shown in Figure 29, where the current source frequency is the same as the mechanical vibration and the magnitude is set by the properties of the piezoelectric material (which determines the capacitance) and the parasitic damping (which determines the amplitude of mechanical motion).

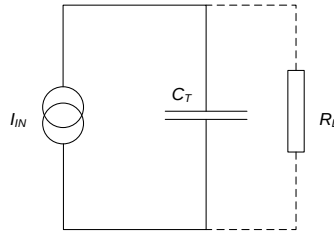


Figure 29 Simplified model of piezoelectric generator (assuming poor electromechanical coupling)

As a consequence, it can be shown [17] that the maximum power that can be dissipated in a linear load resistance (or into an interface circuit with an equivalent input impedance) occurs when the load resistance is given by:

Equation 14
$$R_L = \frac{1}{\omega C_T}$$

It is clear that in this case, the power that can be extracted from the circuit is limited by the intrinsic shunt capacitance of the piezoelectric material. However, if an impedance match as per Equation 14 was presented to the piezoelectric harvester, the mass could potentially hit the end-stops of the harvester. This is because the electrical damping force from an optimal load resistance is not large enough to damp the motion of the mass when the displacement of the harvester is significantly larger than the maximum displacement limit of the proof mass. Unlike the power processing circuits presented earlier in this chapter, a conventional impedance match would not be the best method to use in order to prevent the proof mass from needlessly dissipating energy at the end-stops of the harvester.

Early work on piezoelectric harvesters made use of this resistive match to maximise power output by measuring power dissipated in a simple load resistor [18, 19], although more recent work has attempted to overcome this limitation by using timed switching elements instead of optimised linear resistive loads.

To increase the power output over what can be achieved with a linear resistive load, two steps can be taken:

- Pre-biasing the piezoelectric material before mechanical work is done against it.
- Synchronously extracting charge from the piezoelectric element rather than continuous extraction into a linear resistive circuit.

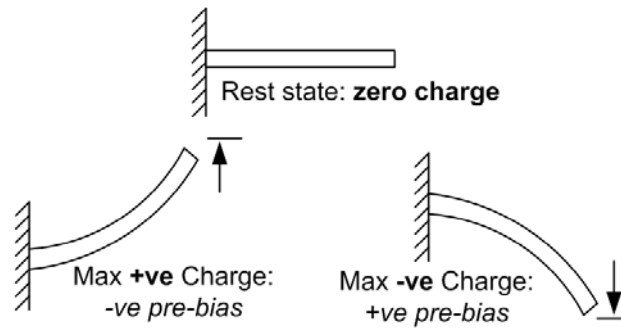


Figure 30 Pre-biasing of piezoelectric increases damping force. (Redrawn from [20])

The first idea, *i.e.* that of pre-biasing, can allow a stronger coupling between the electrical and mechanical systems. The second idea, that of synchronous discharge, overcomes the limitation of real power transfer due to the presence of the intrinsic capacitance. When a piezoelectric material is strained in one direction in open circuit, the resulting charge displacement causes a force which tries to move the material back to an unstrained state, and some work is done in straining the material. If a charge is placed onto the material forcing it to become strained in one direction before the material is forced to move in the other direction by an external force, more mechanical work can be done as the force presented by the piezoelectric material is increased. Therefore more electrical energy can be generated. This is illustrated in Figure 30. When the piezo cantilever is strained upward at maximum displacement such that a positive charge would be generated by the deflection of the material if in open circuit, a negative pre-bias voltage is applied to the material allowing increased mechanical work to be done as the cantilever's free end moves downwards.

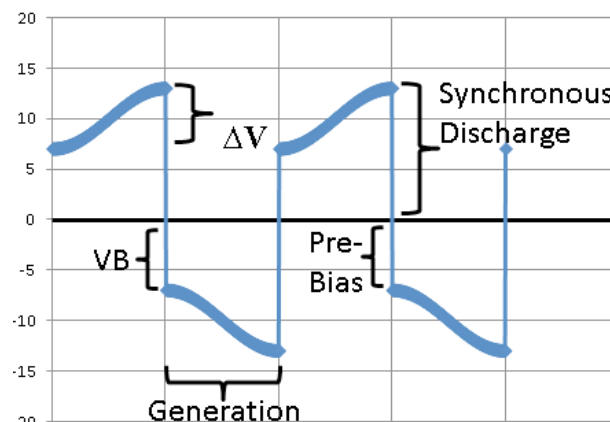


Figure 31 Piezoelectric voltage when operated with pre-bias and synchronous discharge. (Reproduced from [20] with permission from the Transducer Research Foundation)

The opposite applies when the free end of the piezo cantilever is at the maximum downwards position. If the applied bias V_B is large compared to the piezoelectrically induced voltage change ΔV_p , the force

magnitude will now be constant at $\approx \alpha V_B$, rather than oscillating in the range $\pm \alpha \Delta V_p$. The voltage on the piezoelectric material is then as sketched in Figure 31.

The first of these techniques, *i.e.* that of pre-biasing, was originally proposed by Taylor et al. in [21], however Guyomar et al were the first to apply the technique to the low power energy harvesting domain in [22]. An increased power output was demonstrated by inverting the charge from the piezoelectric material at the extremes of the motion. The piezoelectric transducer terminals were also connected to a bridge rectifier and smoothing capacitor, allowing the extraction of power in a useful stable DC form. In [22], the explanation of improved power output is given in terms of the nonlinear functioning of the circuit, but it is the increased mechanical force due to the resultant cell biasing that is the essential origin of the increased output power. The disadvantage of this technique is that the charge extraction from the piezoelectric material cannot be controlled independently of the voltage on the output side storage capacitor. Ultimately this means that the pre-charge bias cannot be optimised for the particular vibration source and mechanical generator characteristics as it is dependent on the storage capacitor voltage and loads resistance. In other words, the optimal electrical damping, detailed in Chapter 5, cannot be set independently of the capacitor voltage.

Their latest results are presented in [23], where they propose a synchronised switch harvesting on inductor circuit with magnetic rectifier (SSHI-MR). This circuit, shown in Figure 32, utilises a transformer with a turns ratio that is much greater than one. The transformer, with two anti-parallel primary windings, allows conversion of the AC piezoelectric voltage to DC. Switches $S1$ and $S1'$ (serially connected to a primary winding) are closed when the displacement of the piezoelectric element reaches its maximum and minimum points respectively. These switches are alternatively opened at half the resonating time period of $\sqrt{LC_0}$, which arises from the series combination of L and C_0 . With the

transformer in place, the threshold at which the diode conducts is lowered to $\frac{V_D}{m}$. This could potentially give a significant reduction in the diode conduction losses when compared with a full diode bridge directly connected to the piezoelectric material. At a displacement amplitude of $23 \mu\text{m}$, vibration frequency of 1 kHz , the SSHI-MR technique resulted in a harvested power of approximately $400 \mu\text{W}$ when an optimal load resistor is used. This harvested power is 56 times greater than when a conventional diode bridge rectifier was used in place of the transformer – signifying the importance of reducing the power losses inherent in discrete power electronics components such as diodes.

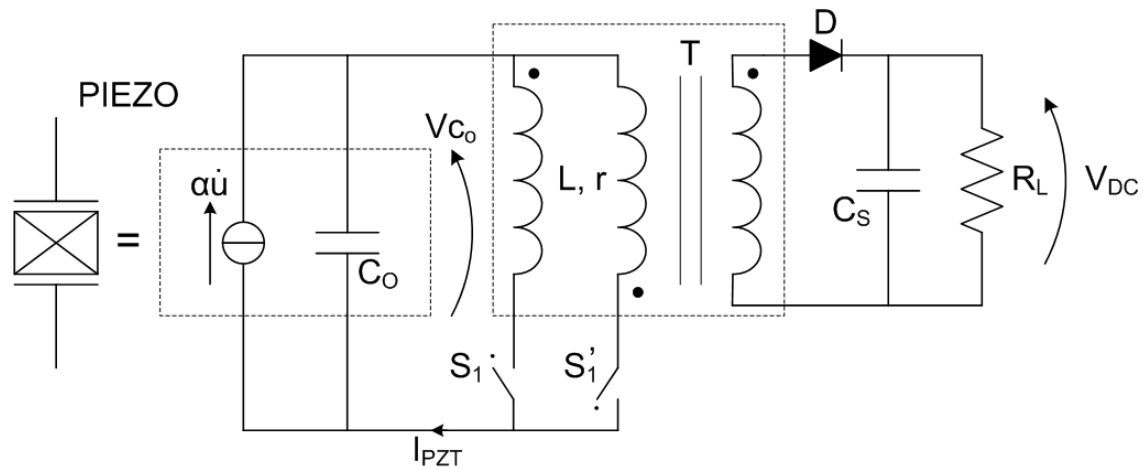


Figure 32 Synchronised switch harvesting on inductor (SSHI) with magnetic rectifier circuit as proposed by Garbuio et al. (Redrawn from [23])

In an attempt to allow optimal pre-biasing without dependence on the status of the load circuit (i.e. capacitor voltage or load resistance), Dicken et al., presented a new approach to increasing the output power from piezoelectric energy harvesters by pre-biasing combined with a synchronous charge extraction circuit.

The key potential improvement of this approach over the techniques presented by Guyomar is that the pre-charge bias circuit and piezoelectric generation cycle can be completely isolated from the output side circuitry and therefore there is no such thing as an optimal load resistance, only an optimal pre-bias voltage. The optimisation of the energy capture by this circuit therefore only depends on the pre-bias voltage applied to the piezoelectric device. The prototyped circuit is shown in Figure 33.

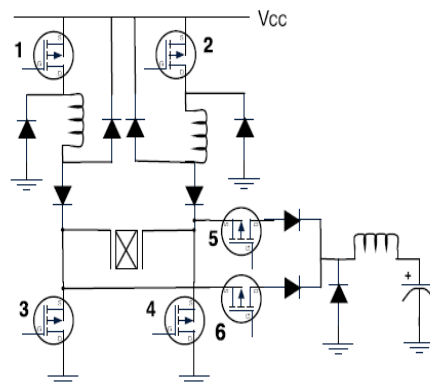


Figure 33 Piezoelectric pre-biasing circuit with synchronous charge extraction. (Reproduced from [20] with permission from the Transducer Research Foundation)

MOSFETs 1 to 4 are used to pre-bias the piezoelectric material at the extremes of the cycle. MOSFETs 5 and 6 are used to extract the energy from the piezo to the output stage just before pre-biasing occurs. Diodes are present to allow recovery of energy stored in inductors to the power supply.

The energy stored in the piezoelectric material's intrinsic capacitance is proportional to the square of the voltage generated by its deflection. If additional charge was added to the piezoelectric material prior to the generation of charge due to mechanical deflections, more work is required to charge the intrinsic capacitance. This is because the voltage on the charge will be higher when compared to the situation where no initial charge was present (no pre-biasing). Once the energy generated from the previous half-cycle of the mechanical deflection is discharged, the piezoelectric material will be pre-biased at its maximum and minimum deflection positions before the material deflects in the opposite direction. To calculate the gain in energy due to the pre-charging condition requires the energy used in charging and discharging the piezoelectric material. Defining the efficiencies of the charging and discharging steps as

η_c and η_d respectively, the energy supplied to charge the piezoelectric material to a voltage, V is $\frac{CV^2}{2\eta_c}$

whilst the useful energy obtained at discharge is $\frac{1}{2}C\eta_d(V + \Delta V)^2$. Variables C and ΔV represent the intrinsic capacitance and the voltage change due to the mechanical deflection of the piezoelectric material. Thus, the net output energy is:

$$\text{Equation 15} \quad E_{out} = \frac{1}{2}C \eta_d (V + \Delta V)^2 - \frac{V^2}{\eta_c}$$

By setting $\frac{dE_{out}}{dV} = 0$, the optimum V in terms of ΔV can be found.

$$\text{Equation 16} \quad V = \frac{\eta_c \eta_d}{1 - \eta_c \eta_d} \Delta V$$

Using Equation 15 and Equation 16, an expression for the optimum energy gain in terms of the efficiency can be obtained. Assuming that $\eta_c = \eta_d = \eta$, the energy gain factor, f_E (ratio of energy generated for synchronous extraction with zero pre-bias to energy generated with the optimal pre-bias for a given efficiency) is:

$$\text{Equation 17} \quad f_E = \frac{E(\eta)}{E(V=0)} = \eta + \frac{3\eta^3}{(1 - \eta^2)}$$

The energy gain factor in Equation 17 is plotted in Figure 34. A high output gain is obtainable at efficiencies greater than 90%.

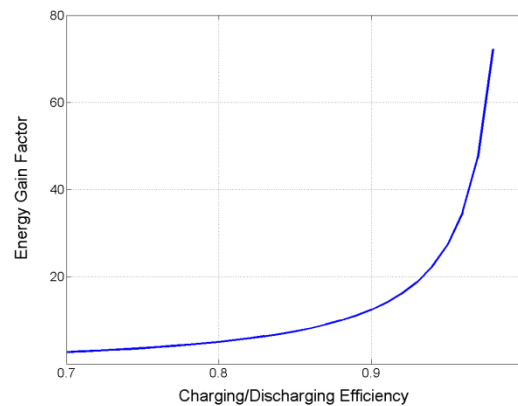


Figure 34 Theoretical power enhancement relative to conventional piezoelectric cell vs. efficiency of pre-biasing [20].

Results presented in [20] showed that the pre-biasing technique produced a net output power of about $110 \mu\text{W}$ at a pre-bias voltage of 12.5 V (Figure 35). This is an increase of approximately 10 times the output power compared to that using a simple optimal load resistance. At the moment, this technique has not shown as much increase in power over a simple optimal resistor as that shown by Guyomar, although in the experimental results shown in Figure 35, breakdown of the semiconductors was the limiting factor.

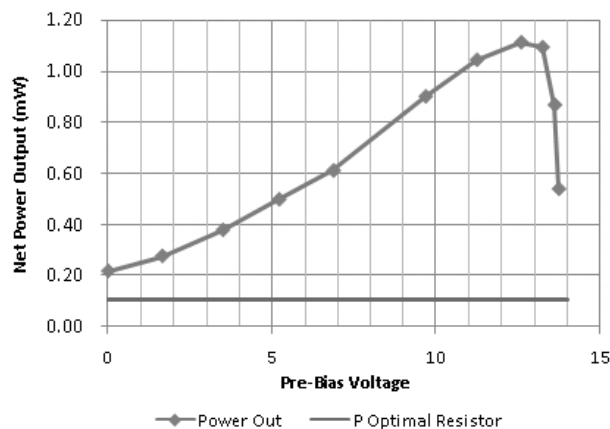


Figure 35 Improvement in net output power with pre-biasing compared to using just an optimal resistive load. (Redrawn from [20])

Electrostatic Harvesters

As discussed earlier in this book, the electrostatic harvester generally uses a moving plate capacitor in order to convert kinetic energy into electrical energy. The existence of the non-constant valued capacitor makes it difficult to model an electrostatic generator using linear circuit components. Only an approximation is possible (Figure 3b) and such an equivalent model does not necessarily give insight into the device operation. It is, however, possible to derive the optimal operation of an electrostatic generator in terms of capacitor voltage and thus to determine the optimal operation of the interface circuitry so as to realise the equivalent of an impedance match for the electrostatic case.

Among all the energy harvesters reported to date, miniaturisation of electrostatic harvesters have been more promising than the other transducer technologies in terms of the creation of true MEMS devices utilising MEMS fabrication techniques at typical MEMS device scales. Consequently, the power electronic circuits presented in this section have generally been designed with a view to the fact that the harvester output powers are very low, in the 1-100 μW range. This minimal power output and the high voltages generated places very difficult constraints on the power electronics in terms of minimising off-state conductance and minimising parasitic capacitance and an example of custom semiconductor device design for an electrostatic harvester is discussed.

There are two main techniques which have been used to realise the electrostatic transducer mechanism. These are switched systems and continuous systems [24], with switched systems being the most studied.

Switched Systems

The switched type of connection between the transducer and the circuitry involves a reconfiguration of the system, through the operation of switches, at different parts of the generation cycle. Switched transducers can further be split into 2 main types:

- Constant charge
- Constant voltage

When the transducer is operated under constant charge, the plates are separated away from one another with a fixed overlap area. However, under constant voltage operation, the plates are moved relative to one another while maintaining a fixed gap between them. The conditions that the interface electronics must present to the harvester in order to extract power optimally can be found using the forces present on the plates of the capacitor as shown in Figure 36. The rate of change of capacitance with respect to distance differs depending on the axis of the relative motion of the two plates: x_{perp} for perpendicular motions and x_{par} for parallel motions. Consequently, for a given electric field strength and plate area, the force between the plates not only depends on the distance between the plates, but also on the axis of relative motion. These forces are indicated as F_{perp} (perpendicular force) and F_{par} (parallel force) in Figure 36. Using the principle of virtual work, the perpendicular and parallel forces acting on the capacitor plates can be found, depending on whether the charge or the voltage across the capacitor plates is held constant.

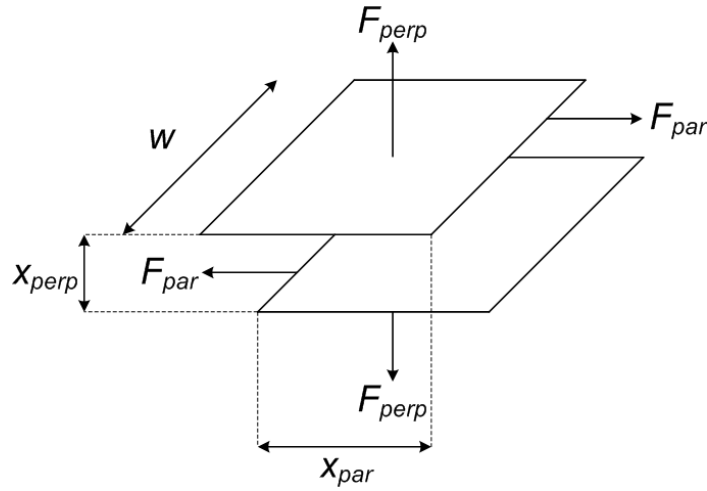


Figure 36 Forces acting on charged capacitor plates.

Perpendicular Force

The energy stored in the parallel plate capacitor in Figure 36 is:

Equation 18
$$Energy = \frac{1}{2} \frac{Q^2}{C} = \frac{1}{2} Q^2 \frac{x_{perp}}{\epsilon w x_{par}}$$

When the plates of the capacitor experience a change in the perpendicular direction (x_{perp}) with the plates having a fixed amount of charge, work is done against the electric field between the plates and electrical energy will be generated. As the plate separation increases, additional potential energy is stored in the increased volume of electric field. The perpendicular force acting on the plates can be found by differentiating the equation for energy with respect to the perpendicular separation of the plates (x_{perp}).

Equation 19
$$F_{perp} = \frac{1}{2} \frac{Q^2}{\epsilon w x_{par}}$$

Parallel Sliding Force

Moving the relative positions of the plates such that the overlapping area between them varies with time will change the capacitance between the electrodes. Using the principle of virtual work,

Equation 20
$$Energy = \frac{1}{2} V^2 \frac{\epsilon w x_{par}}{x_{perp}}$$

If the capacitor plates have a fixed voltage across them and are moved relative to one another but with a constant separation distance, the electric field strength remains constant but current is forced to flow because the volume of the electric field decreases.

Equation 21
$$F_{par} = \frac{1}{2} V^2 \frac{\epsilon w}{x_{perp}}$$

The expressions of perpendicular and parallel forces acting on the plates of the electrostatic harvester provides an indication of how much electrical damping should be applied to the harvester for it to operate optimally and not hit the end-stops. For both the constant charge and voltage cases, the optimal electrical damping force that results in an optimised power output from the harvester is given by:

Equation 22
$$F_{opt} = \frac{Y_0 m \omega^2}{\sqrt{2}} \frac{\omega_c}{|(1 - \omega_c^2)U|}$$

Y_0 is the displacement of the electrostatic harvester, m is the proof mass, ω is the frequency of vibration, ω_c is the frequency of vibration normalised to the resonant frequency of the harvester. The variable U is defined as:

Equation 23
$$U = \frac{\sin \frac{\pi}{\omega_c}}{1 + \cos \frac{\pi}{\omega_c}}$$

Each variable in Equation 22 and Equation 23 has a specific value depending on the operation of the electrostatic harvester. So, to extract power optimally from the harvester, the interface electronics has to provide an electrical damping force equivalent to Equation 22 by delivering the correct amount of charge or voltage to the transducer. This is equivalent to an impedance match for the electrostatic case. If the applied electrical damping force is greater than the sum of the inertial and spring force (harvester modelled as a mass-spring-damper system), the mass will cease to move relative to the harvester's frame and no energy is generated.

Examples of Interface Electronics for Constant Charge Operation

This type of electrostatic harvester operation was reported in [25] for a MEMS fabricated energy harvester. The prototype was fabricated using techniques such as DRIE and the movable capacitor plate had an active area of approximately 200 mm². In Figure 37, at around 50 ms, the capacitor is pre-charged, at maximum capacitance, to around 30 V. After some time, the source motion causes the plates to separate. This operation is done under constant charge and so a large increase in voltage can be seen. Once the electrodes reach maximum separation, the capacitor is discharged. This generator was shown to generate around 12 µJ from an input motion of 40 Hz and 6 mm amplitude.

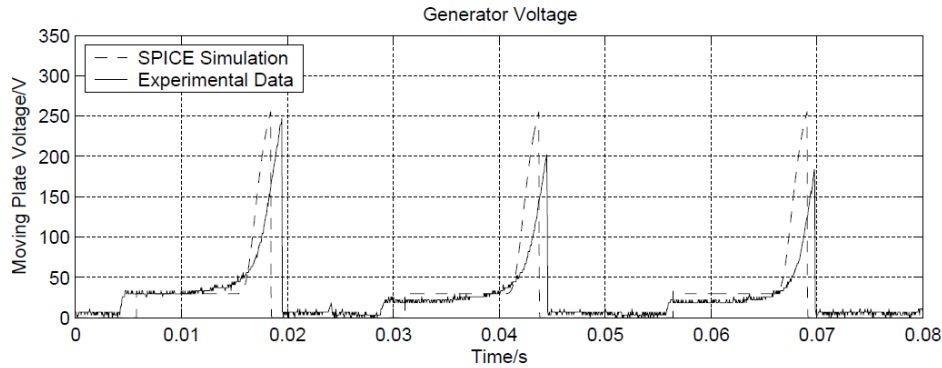


Figure 37 Simulated and experimental data for an electrostatic generator operating under constant charge from [25].

A suitable power conversion circuit for the output side of the aforementioned generator in [25] is the half-bridge step-down circuit shown in Figure 38. The half-bridge has been chosen so that a boot-strap drive can be used to turn on the high-side semiconductor switch, in this case a MOSFET. Although the generation cycle time is long (circa 10 ms) and unpredictable, the power converter need only operate for less than 1 ms to completely discharge the capacitor and so the boot-strap technique is viable. It is desirable to use an integrated inductor, and inductance values in the range 1–10 μH appear to be achievable [26]. The discharge of the generator will occur in a short current pulse and controlling this current through chopping would require a high switching frequency and consequently the associated power losses will be undesirable.

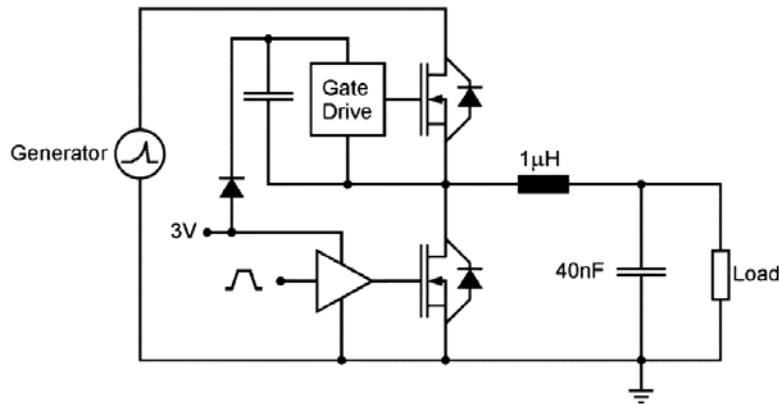


Figure 38 Half bridge converter proposed in [27], the low side MOSFET is only required for boost-strap gate-drive.

It is convenient to split the operation of the circuit in Figure 38 into three phases, as shown in Figure 39. The converter is used in single-pulse mode and the source is weak enough to be completely discharged within a few nanoseconds. In the first phase, during the turn-on of the MOSFET, current flows into the diode to establish a reverse bias and to allow the voltage over the MOSFET to reduce. This current is supplied by the generator and this is an unwanted loss of charge. During the second phase, the inductor current increases and the generator voltage falls until the generator is completely discharged. At this point the inductor current is at its maximum. Then the longest phase begins in which the current free-wheels through the diode until the inductor is demagnetized.

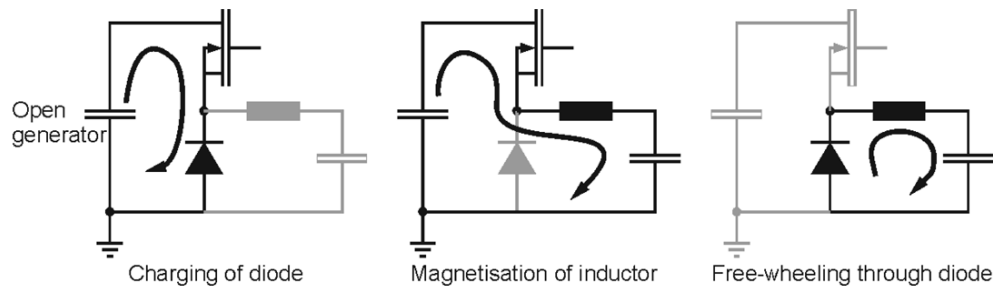


Figure 39 Three phases of conversion with distinct current patterns.

As a first step to designing the circuit in Figure 38 for the MEMS harvester, an assessment was made of the input resistance and capacitance that the circuit must present in the off-state at the maximum generator voltage in order not to compromise generation. The generator's electro-mechanical system was simulated numerically using Matlab for a range of static impedances on the generator outputs, assuming a 20 ms flight time. The requirements are unusually strict: to maintain 80% of the generated energy the off-state loading should be more than $10^{12} \Omega$ and less than 1 pF [4]. These values are not available with standard discrete MOSFETs rated for 300-V blocking. By assuming that the parasitic components of the converter are constant, their effect on the energy generation is analysed and plotted in Figure 40.

To achieve this high level of impedance, thin layer silicon on insulator technology based semiconductors must be designed. In [27], in depth simulation studies were carried out to optimize the MOSFET and diode device areas to optimise the energy generated from the system, taking into account conduction loss and charge sharing effects. A cross-section through the custom MOSFET is shown in Figure 41. It was found that the on-state voltage drop of the MOSFET predominantly affects the conversion efficiency because of high peak currents, which are due to the low inductance used in the circuit in order that the inductor could be integratable on chip.

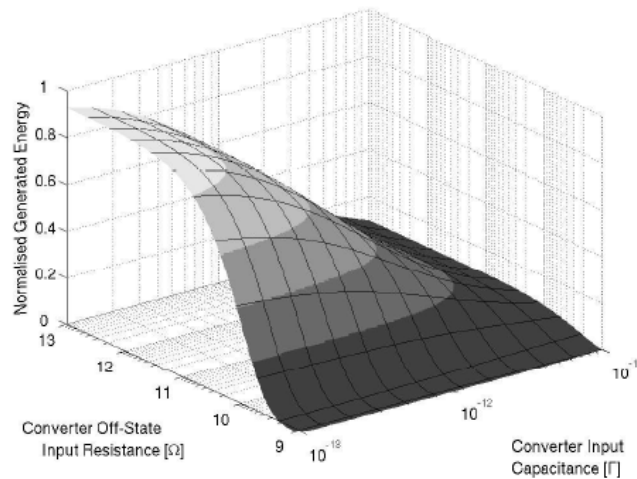


Figure 40 Dependence of generated energy on converter impedance

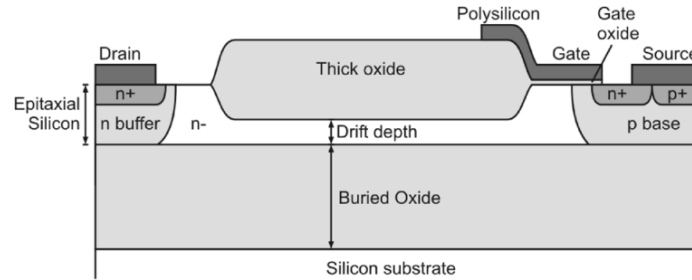


Figure 41 Custom designed silicon on insulator MOSFET for MEMS electrostatic harvester from [27]

In the above example, interface electronics would be required to charge the variable capacitor through an external pre-charge power supply (probably battery) at maximum capacitance and to discharge the variable capacitor through a load (or to recharge the battery) at minimum capacitance. The discharge circuitry alone is not sufficient to make a working energy harvester system. An example of a more complete system with both input and output side electronics for the electrostatic transducer is shown in Figure 42. A charge pump circuit is used to charge and discharge the variable capacitor. Diode $D1$ will be on when the variable capacitor is at minimum position i.e. capacitance is maximum. Diode $D2$ will be on when the voltage at node A is more than the load voltage. Both the diodes will be off during rest of the vibration cycle period. Diodes with low reverse leakage current are suitable for this application to reduce the leakage power loss. JFETs working in a diode mode have been used in [28] to reduce the reverse leakage current.

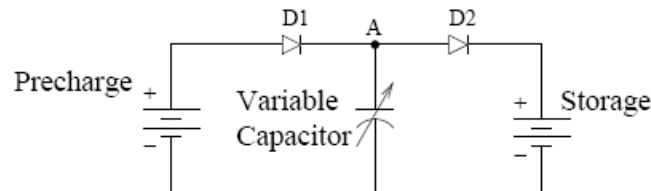


Figure 42 Basic charge pump circuit

The basic circuit of Figure 42 will eventually discharge the energy in the pre-charge source and to avoid this, a flyback inductor was used as shown in Figure 43. Charging and discharging of the variable capacitor is done using the charge pump circuit and the flyback inductor was used to transfer the energy from the temporary storage capacitor (C_{store}). Energy will be stored in the inductor by turning on the MOSFET and when the MOSFET is turned off, the inductor current will free wheel through diode D_{FLY} . The MOSFET gate pulse need not be synchronised with the vibration cycle, which is the case of modified charge pump circuit hence, reducing the complexity of the circuit. Detailed analysis of calculating the efficiency of power conversion is given in [29].

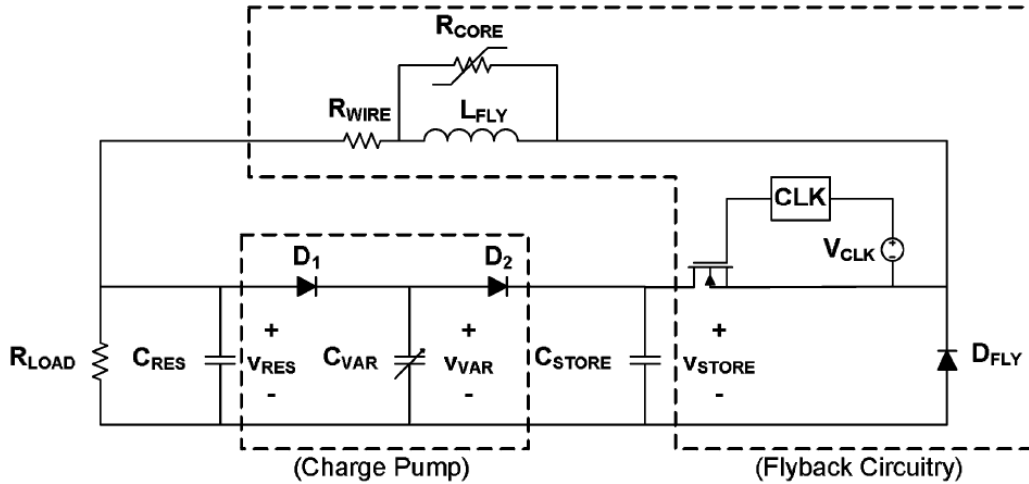
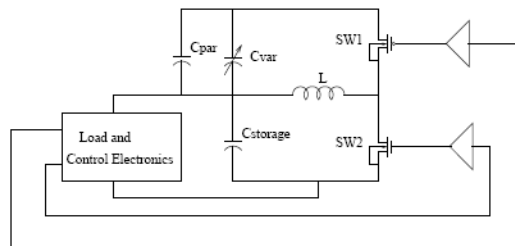


Figure 43 Capacitive energy harvester with source-referenced clock controlling the flyback switch [29].

Examples of Interface Electronics for Constant Voltage Operation

To further reduce the losses such as forward conduction losses, active switches are used instead of diodes in [30] and the modified charge pump circuit is shown in Figure 44. Energy conversion from the mechanical to electrical domain was implemented using low-power digital control circuitry consisting of a delay-locked-loop (DLL) capable of synchronising the energy extraction mechanism to the source vibration frequency (ω in Equation 22). Upon achieving this phase-lock, the reference clock in the digital circuitry will be in-phase with the motion of the generator's moving plate. This enables the generation of the timing pulses for the gates of $SW1$ and $SW2$. During the precharge condition, $SW2$ will be switched on to store energy in inductor L . The stored inductor energy will be used to charge the variable capacitor C_{var} by turning on and off $SW1$ and $SW2$ respectively. During the discharge period, the opposite switching sequence of the pre-charging condition will be implemented to discharge C_{var} . Simulation results of the digital control circuit in HSPICE predicted a control overhead of around $3 \mu W$. The electrostatic generator was predicted to produce $8.6 \mu W$ of power, leaving $5.6 \mu W$ of electrical power for the load electronics.



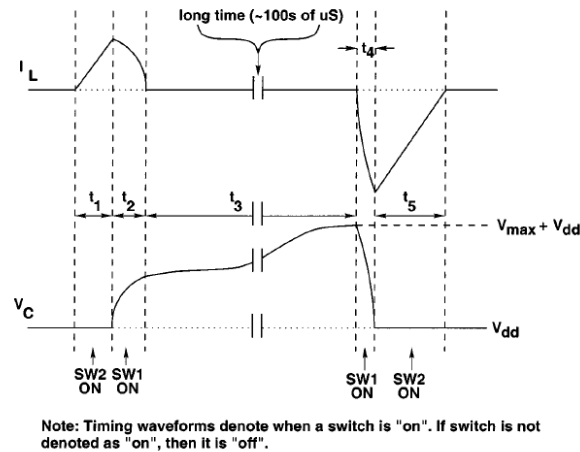


Figure 44 Modified basic charge pump circuit (top) and waveforms (bottom) [30].

Another example of a power processing circuit for a voltage constrained electrostatic microgenerator is shown in Figure 45. During the pre-charge condition, SW2 and SW5 will be switched on to store energy in the inductor L . Switches SW3 and SW4 will be turned on by simultaneously turning off SW2 and SW5 to charge the variable capacitor C_{var} . The unidirectional switch SW1 will be turned on to allow the current to flow from variable capacitor C_{var} to the battery. When the variable capacitor has reached its minimum value, SW1 will be turned off. In order to completely recover the charge across the variable capacitor, reverse switching sequence of the pre-charge condition is used. A complete description of the circuit with waveforms has been discussed in [31].

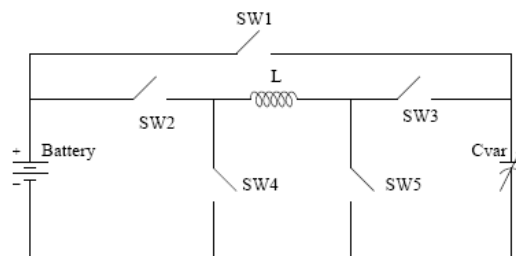


Figure 45 Constant voltage based electrostatic microgenerator for battery charging applications [31].

Continuous Systems

A third mode of operation exists when the variable capacitor is continuously connected to the load circuitry, and this load circuitry provides the capacitor with a polarisation voltage. A simple example of this is a voltage source, a resistor and a variable capacitor wired in series. A change in capacitance will always result in a charge transfer in between the electrodes through the load resistance causing work to be done in the load.

The switched generators previously discussed are special cases of this continuous mode generator: a constant charge generator is equivalent to a continuous generator operated with infinitely high load

impedance, whilst the constant voltage generator corresponds to a continuous generator which is short circuited. Because no work can be done when either the generated current or the generated voltage is zero, these extremes of operation require a switching circuit to make them operate. The use of controlled switches complicates the implementation of the generator and the circuitry required to control them consumes a minimum amount of the generated power and so in some circumstances the use of a continuous system is preferred.

Electrets are often used in combination with a variable capacitance to make a continuous mode generator. The fixed charges of the electret induce an electric field between the electrodes of the capacitor, corresponding to a potential of several tens of volts. Three possible QV diagrams showing the operation of a continuous electret generator are shown in Figure 46a. If the capacitor is operated in a constant voltage mode, a change in capacitance will result in a current through the load circuitry along curve (1-3-1). A high impedance load forces the generator to operate in constant charge as the high impedance obstructs the charge transport between the electrodes (1-2-1). In both of these cases, the area of the QV loop integral is zero as the transition from maximum to minimum capacitance occurs on the same trajectory. An optimised load for a continuous generator will operate the generator in between these extremes along (1-4-1), and as can be seen, work is now done and the loop integral has a finite value. This class of generators is referred to as velocity damped generators because the damping force is approximately proportional to the relative velocity between the proof mass and the frame.

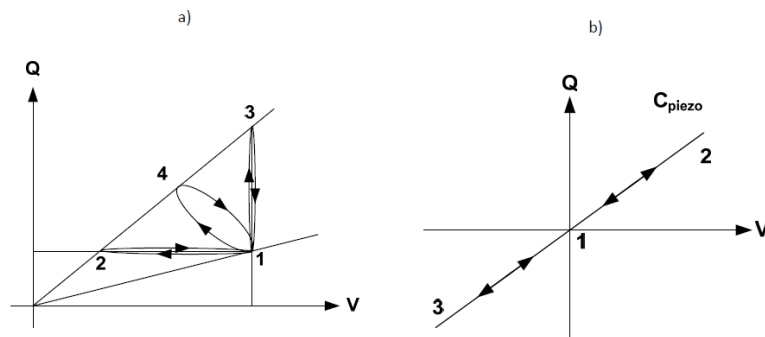


Figure 46 Operation of an electrostatic generator in continuous mode (a) or using piezoelectric polarisation (b).

Examples of Interface Electronics for Continuous Mode Operation

Sterken *et al.* have micromachined a prototype of a 0.1 cm^2 electrostatic micro-generator using silicon-on-insulator (SOI) methods. This comb-like structure was predicted to be able to generate $50 \text{ }\mu\text{W}$. An electret precharges the moving plate of the capacitor (up to a limit of 50 V to prevent clamp-down) which is suspended by meandered beams that function as springs. The lateral displacement of the moving plate changes its capacitance and charge thus, causing a current to flow through the load resistor, R as indicated on the left side of the top diagram in Figure 47. This load resistor is representative of prospective power management electronics to condition the power from the electrostatic generator.

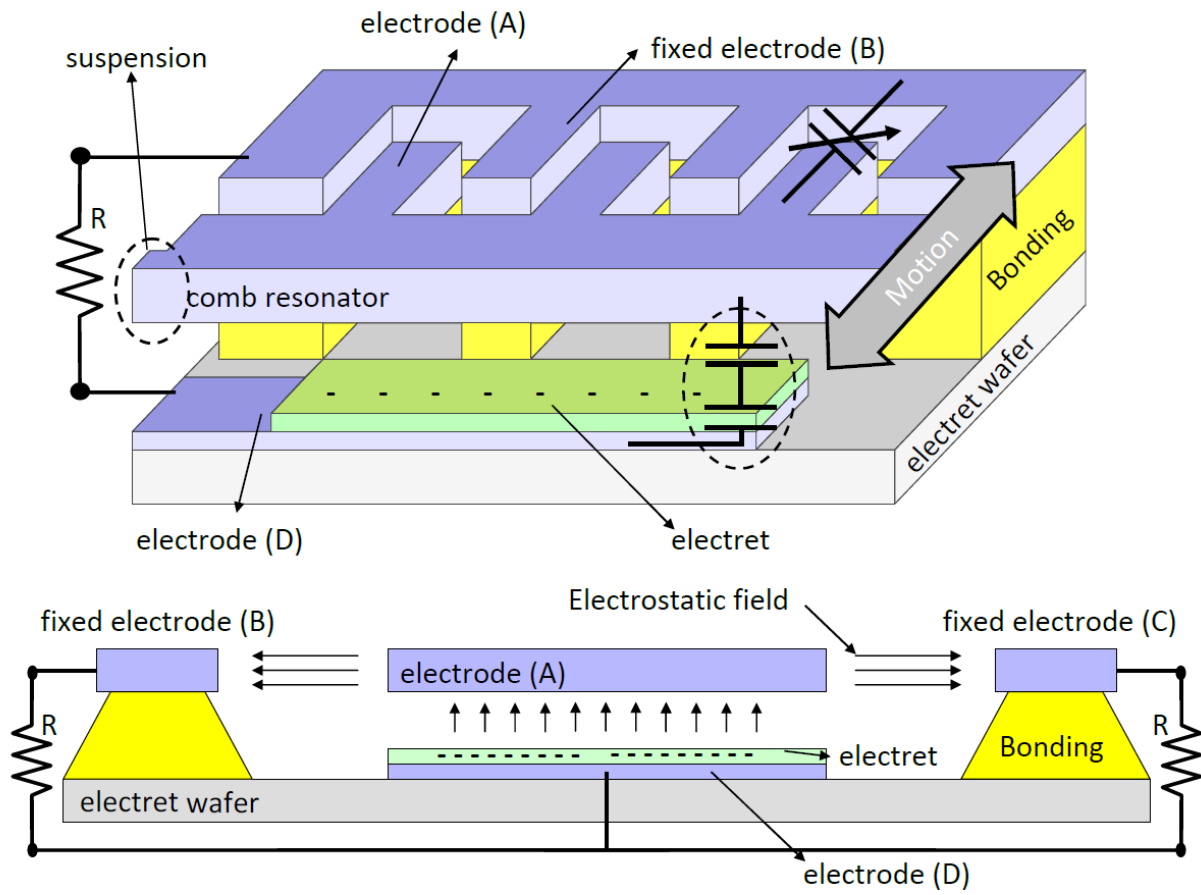


Figure 47 Cross-section (top) and side-view (bottom) of a micromachined electrostatic generator. (Redrawn from [32])

Interface Circuits for Thermal and Solar Harvesters

Now that the electronics for motion-driven harvesters has been described in some detail, we turn our attention to interface circuits for non-kinetic energy harvesters, namely thermo electric generators and solar cells. As previously, we must first determine a suitable model of the source to which our electronic interface must connect. The main difference between these harvesting methods and the kinetic devices is that there is almost no frequency dependence in these models. As such, the dynamics of the energy source can effectively be ignored and the system can be analysed at DC.

Thermal

A structure of a thermoelectric generator (TEG) is shown in Figure 48 [33]. The thermoelectric circuit is formed by using two types of semiconductor material, p-type and n-type, which are connected electrically in series and thermally in parallel. A ceramic plate (electrically insulating but thermally conducting) forms a connection between the heat source (heat sink) and the hot-side (cold-side) of the thermocouple. The rate of heat exchange is denoted by Q_H and Q_C where the subscripts represent hot and cold temperatures respectively. ΔT_{TEG} is the temperature difference between the hot (T_{HJ}) and cold junctions (T_{CJ}) of the TEG whereas ΔT is the temperature gradient on the exterior of the generator.

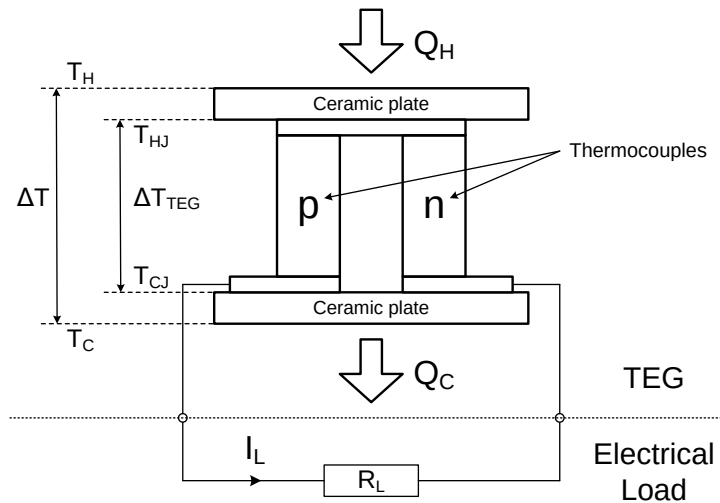


Figure 48 Structure of a thermoelectric generator (TEG) connected to a load resistor (R_L). (Redrawn from [33])

The thermoelectric effect resulting from a temperature difference between two conductors depends on the Seebeck coefficient of the two materials (α_p and α_n). Equation 24 defines the open-circuit voltage generated from a TEG.

Equation 24
$$V_G = \alpha_{pn} \cdot (T_{HJ} - T_{CJ})$$

When a load is connected to the TEG, a current I_L flows as per Equation 25. R_{int} is the internal electrical series resistance of the TEG.

Equation 25
$$I_L = \frac{V_G}{R_{int} + R_L}$$

R_{int} is a function of the height (h) and cross-sectional area (A) of the thermocouple and the resistivity (ρ) of the material.

Equation 26
$$R_{int} = \frac{2\rho h}{A}$$

Figure 49 shows the electrical equivalent circuit of the TEG connected to a resistive load based on Equation 24 and Equation 25. This seems to be a very simple model of the device and shows that the main requirement for the interface circuitry is to set its input impedance to the electrical resistance of the TEG. However, before making this simple conclusion, we must first determine any dependence that a load current may have on the temperature across the device and therefore on the thermoelectric voltage, V_G .

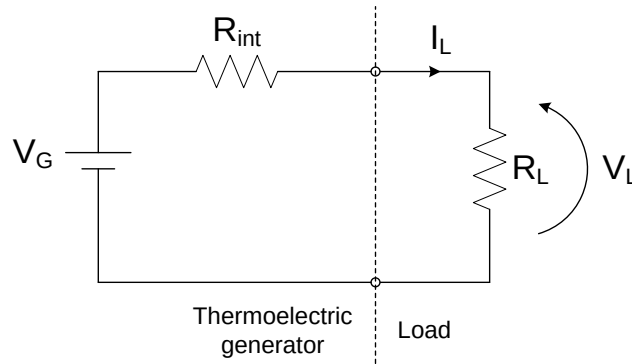


Figure 49 Electrical equivalent circuit of the thermoelectric generator connected to a resistive load.

With reference to Figure 49, the rate of heat exchange between the hot and cold junctions of the TEG, i.e. Q_H and Q_C is given by Equation 27 and Equation 28. K is the thermal conductance of the ceramic plates.

Equation 27
$$Q_H = K \cdot (T_H - T_{HJ})$$

Equation 28
$$Q_C = K \cdot (T_{CJ} - T_C)$$

Both Q_H and Q_C can be described as the sum of the Peltier effect, thermal conduction through the p- and n-thermocouples and the heat loss in the internal series resistance of the TEG as:

Equation 29
$$Q_H = \alpha_{pn} T_{HJ} I_L + K_{int} \cdot (T_{HJ} - T_{CJ}) - \frac{1}{2} I_L^2 R_{int}$$

Equation 30
$$Q_C = \alpha_{pn} T_{CJ} I_L + K_{int} \cdot (T_{HJ} - T_{CJ}) + \frac{1}{2} I_L^2 R_{int}$$

In Equation 29 and Equation 30, K_{int} is the internal thermal conductance of the thermocouples and can be expressed as Equation 31 where λ is the thermal conductivity of the thermocouple material.

$$\text{Equation 31} \quad K_{int} = \frac{2\lambda A}{h}$$

Equating Equation 27 to Equation 29 and Equation 28 to Equation 30, the effective temperature gradient across the hot and cold junctions, ΔT_{TEG} , can be found.

$$\text{Equation 32} \quad \Delta T_{TEG} = \Delta T \frac{K}{K + 2K_{int} + \frac{\alpha_{pn}^2 (T_H + T_C)}{R_{int} + R_L}}$$

The third term in the denominator which involves both R_{int} and R_L is a consequence of the Peltier effect of the load resistance. If the load resistance was reduced, the current flowing from the TEG (I_L) will increase. As a result, the hot junction of the TEG will experience a loss in heat and the cold junction will become hotter due to the Peltier effect. In general, ΔT_{TEG} is affected whenever R_L is changed. However, the term $\frac{\alpha_{pn}^2 (T_H + T_C)}{R_{int} + R_L}$ is typically small (α is within the mV/K range) compared to the other terms in the denominator and is usually neglected. This greatly simplifies the source model of the TEG.

The output power (P_{out}) from the TEG is then,

$$\text{Equation 33} \quad P_{out} = I_L \cdot (\alpha_{pn} \Delta T_{TEG} - I_L R_{int})$$

Substituting Equation 25 into Equation 33 gives

$$\text{Equation 34} \quad P_{out} = (\alpha_{pn} \Delta T_{TEG})^2 \frac{R_L}{(R_{int} + R_L)^2}$$

Clearly in this case, P_{out} is maximized when R_L is matched to R_{int} , assuming the Peltier effect that the load resistance has on the TEG is negligible. Therefore, the optimal value of R_L is simply equal to the measured electrical resistance between the terminals of the TEG.

It is possible to increase the thermoelectrically generated voltage by connecting N -thermocouples electrically in series (thermally in parallel). V_G , R_{in} and K_{int} will increase proportionally to N whereas the output power increases by N^2 . Intuitively, cascading multiple thermocouples seems like an attractive solution to overcome the low voltage levels from TEGs. Limitations of size are generally important in wireless sensor nodes and this constrains the total number of thermocouples that can be used in a TEG.

Example Interface Circuits for Thermoelectric Generators

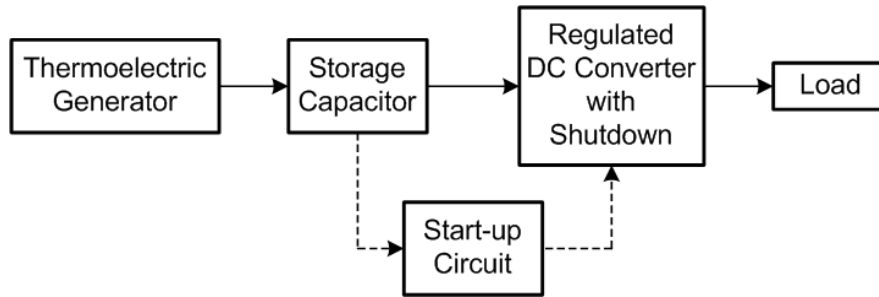


Figure 50 Block diagram of a generic power processing setup for the thermoelectric generator.

As an example, consider the work reported in [34] where approximately 100 μW of usable electrical power was extracted from a TEG attached to the human body when the ambient temperature is 22°. The reported open circuit voltage of the TEG under matched electrical loads was between 0.6 – 1.0 V. These low voltage levels are generally insufficient to power a sensor or load electronics. In addition to that, if the ambient temperature changes, the voltage levels from the TEG vary and consequently some form of voltage regulation is needed to supply a load with a constant voltage. By choosing a regulated DC converter with a shutdown input (Figure 50), a start-up circuit can be used to keep the converter in shutdown mode until the storage capacitor has accumulated sufficient charge to overcome the minimum input voltage of the converter. When this happens, the start-up circuit will disable the shutdown mode and the converter can begin regulating the voltage across the load. Moreover, the storage capacitor can discharge in order to supply additional power to the converter in the event of a surge in the load current.

Mateu et al. reported on a duty-cycle controlled maximum power point tracker circuit that was designed for use on a TEG [35]. Due to the low voltage levels from the TEG, a boost converter was chosen to perform the step-up conversion on the voltage. The maximum power point was tracked by changing the duty cycle of the converter such that the output voltage from the TEG was half that of the generated voltage from the TEG. Under such circumstances, the load as seen by the TEG would be equal to that of the TEG's internal series resistance. Hence, an impedance match is present between the TEG's output terminals and its immediate interface, the boost converter.

Power Electronics for Photovoltaics

At the scale of renewable power generation, a typical configuration of the power processing circuits for photovoltaic cells consists of a boost converter interfacing to a DC link and an inverter making a grid connection. In an energy harvesting device the interface circuit should have the same configuration and function as the boost converter and of course the inverter for grid connection is not required.

Maximum Power Point Tracking (MPPT) for PV Arrays

For a set level of illumination, photovoltaic (PV) arrays generate different amounts of electrical energy depending on the current being drawn from them. Typical voltage/current profiles from a PV cell are shown in Figure 51, where each curve represents a particular combination of light intensity and temperature. Clearly, maximum power is extracted from the cell when product of cell current and voltage is maximised. For each curve, these points are labelled on as P_{MPP} .

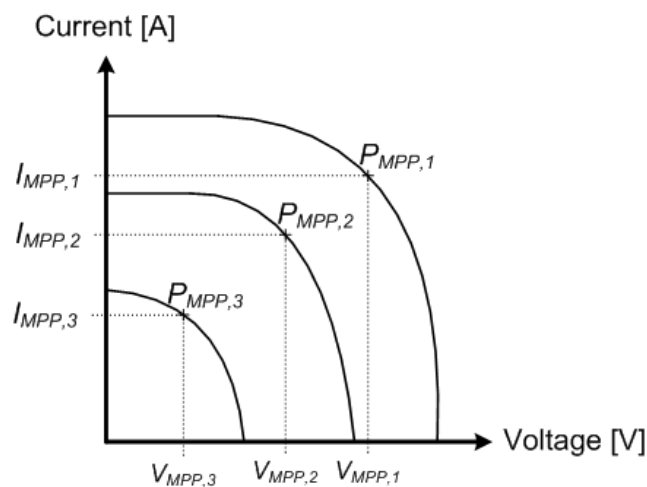


Figure 51 Characteristic PV array I – V curve at different operating conditions.

Therefore, as has been the case with other energy harvesting methods, in order to extract maximum energy from the transducer, an optimal load resistance must be connected in the form of a power converter emulating that resistive load as its input impedance. Setting the load impedance is relatively simple as long as the optimal load is known. The difficulty with PV, and indeed perhaps all harvester technologies, is that finding the optimal load as operating conditions change, can be difficult. This difficulty arises from the fact that the maximum power point is dependent on the temperature, the irradiation level and the age of the solar cell. Some estimation could be made by measuring temperature and irradiation level but the usual method employed in large scale PV installations is to continually hunt for the maximum power point, modifying the duty cycle of a power converter to ensure maximum power is extracted. The Perturb and Observe (P&O) method has to date been the preferred technique for hunting for the MPP. The duty cycle of a power converter is continually perturbed and if a

change in duty cycle increases the power output, the duty cycle is again changed in the same direction. If the change in duty cycle produces a reduction in output power, the duty cycle is next perturbed in the opposite way and as such, the maximum power point of the solar array is attained.

Figure 52 shows a typical output power vs. output voltage curve of a PV array when operated under static conditions. If the initial output voltage of the array results in an output power at point A, an increment of Δ_1 will move the operating point to the MPP position. However, at point B, decreasing the output voltage by Δ_2 will result in optimal operation of the PV array.

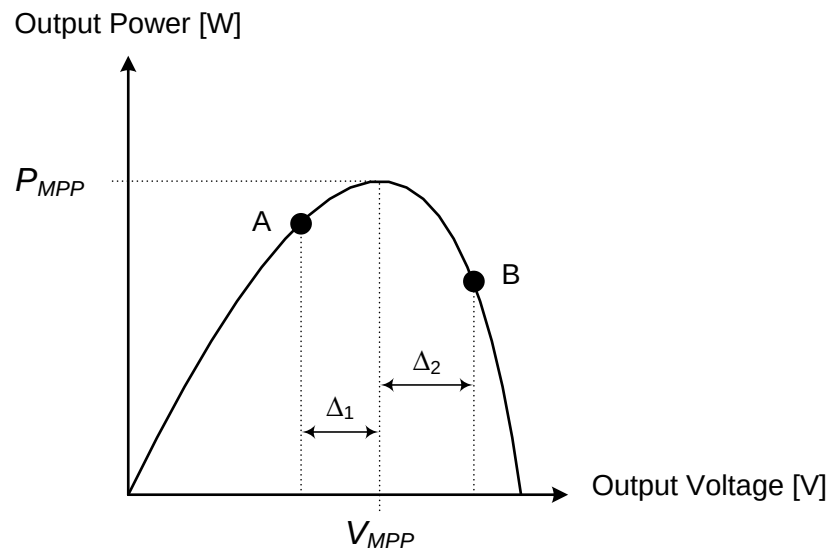


Figure 52 Output power vs. output voltage characteristic of a PV array when operated under static conditions.

Power Electronics for Energy Harvesting PVs

At the time of writing, little attention has been given to MPPT control of the interface circuitry for energy-harvesting sized PVs, although miniature boost DC-DC converters, suitable for interfacing to PVs have been designed, such as [36]. In such cases, the main difficulty comes from the very low input voltage produced from individual solar cells. In [36], the authors have considered the issue of low voltage start up; it is not possible to run the converter's control circuitry from a few hundred mV available from the energy harvesting transducer, but it is possible to draw energy from the supply, given that the control circuit for the power converter is operational. In order to achieve this, the authors used a secondary winding on the input inductor with a resonant capacitor to drive the gate of a JFET. Once the circuit has started up, the JFET is no longer used and a parallel MOSFET allows normal controlled operation.

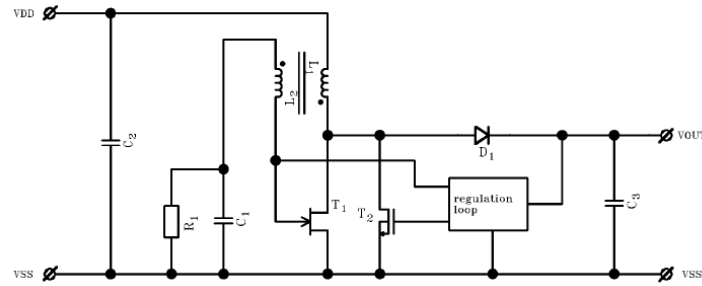


Figure 53 Low voltage start-up DC-DC converter for thermoelectric or solar harvesting applications (Reproduced from [36] with permission from the Transducer Research Foundation)

As stated previously, very little attention has been given on the energy harvesting scale to maximum power point tracking control for low power solar harvesting. Figure 54 shows the configuration of a typical larger solar cell arrangement with MPPT tracking. Here, the MPPT circuit generates the pulse-width-modulated (PWM) signal that drives the switching transistor in the boost converter. The duty cycle of the PWM signal will depend on the real-time values of the PV array's output voltage and current (V_{PV} and I_{PV}). In this example, the boost converter is the immediate interface between the PV array and the load. This is one of the preferred interfaces to a PV array because the boost converter is able to step-up the output voltage from the PV array. Furthermore, the combination of inductor and output capacitor of the boost converter has a smoothing effect on the output current, resulting in smaller output voltage ripples relative to the average output voltage.

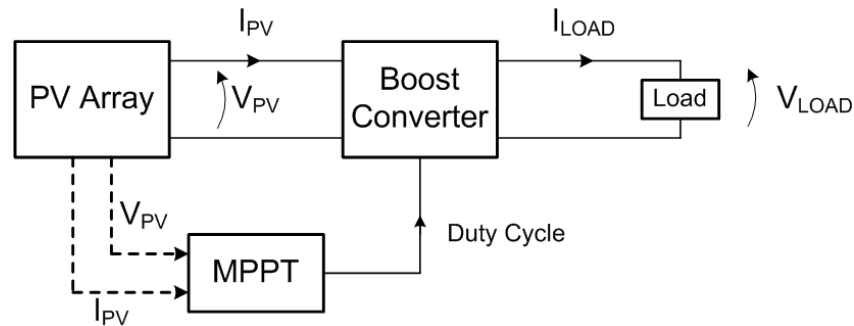


Figure 54 A simplified schematic showing the power processing circuitry.

The constraints of implementing such MPPT circuitry on the scale of an energy harvesting device is simply that when taking into account the power consumption overhead of the control circuitry, the net power output from the system should be greater than the output where MPPT is not applied. Therefore, the success of such circuitry is dependent on both designing a low power MPPT implementation and the benefit in increased power extraction from the PV cell, which is highly dependent on the degree of variability of light levels irradiating the device.

Energy Storage Interfaces

The vast majority of energy harvesting transducers will not be able to supply energy at a constant rate over long periods of time. Clearly, a solar cell can only produce electrical energy when illuminated and a vibration harvester can only produce electrical energy when it is subjected to acceleration. However, many applications of energy harvesting technology may require a constant source of electrical energy to supply the load. Clearly, if the average power consumption of the load is greater than the average power generated by the harvester, it is not possible to provide power continually to the load. However, if the average power generated is equal to or exceeds average consumption by the load, it is possible to run the load continually. When excess power is harvested, it is stored in the storage component and when there is insufficient power from the harvester, the storage component can be discharged to supply the load electronics. Besides that, the energy storage component is capable of handling surges in load currents during events like a turn-on transient of the load electronics. As discussed in Chapter 3, electrical storage in the form of a battery or capacitor is generally used.

Using a super-capacitor, such as those shown in Figure 55, for the storage element has the advantage that pushing energy into it is a relatively simple task with few constraints on how the capacitor is charged. In the systems described previously in this chapter, the interface circuit between the energy harvesting transducer and the storage capacitor only needed to have a controlled input; the interaction between that circuit and the storage capacitor was ignored. In other words, we were free to alter the operating mode of the interface circuit to optimise the operation of the transducer without taking into account how this affected the storage element. Fundamentally this is because a storage capacitor is very tolerant to the rate at which energy is transferred into and out of it. There is of course one constraint that must be taken into account for interfacing with a storage capacitor; its breakdown voltage. When the capacitor has reached its maximum voltage, the interface circuit must stop transferring energy from transducer to storage to prevent breakdown, *i.e.* harvesting must stop. However, whilst the use of a storage capacitor makes the design of the interface circuit simpler, the disadvantage of using capacitive storage is the wide voltage range that it operates over. This in turn means greater difficulty is encountered in regulating the harvester system's output voltage for the load electronics, as a very wide-input power converter is required.



Figure 55 Super-capacitors from AVX [37].

When using batteries as the energy storage element, the opposite is true. The rate and way in which the battery is charged can significantly influence the lifetime of the cells. However, as the battery voltage is relatively constant, load voltage regulation is much simpler than in the capacitive storage case and, indeed, output regulation may not even be required as long as the cells are carefully chosen. Limits such as available cell voltages must be taken into account. As an example, lithium-ion cells have nominal voltages of around 3.7V [38] and therefore it is not possible to power lower voltage circuitry from Li-ion cells without some form of output voltage regulation.

Output Voltage Regulation

As we have discussed, fluctuations in the voltage across the energy storage component means that the system may require some form of load voltage regulation. The fluctuation may be negligible if a battery is used, but may be significant if a capacitor is used as the storage component.

Output voltage regulation can in some cases be achieved by using off the shelf linear or switching voltage converters however the inefficiency of a linear regulator makes them unsuitable for wide input fixed output conversion. Therefore, when using capacitive storage, a wide-input switching regulator is the preferred interface between the energy storage component and the load electronics. Presently, commercial off-the-shelf switching regulators at the tens to hundreds of mW capability have reported efficiencies of around 90% [39].

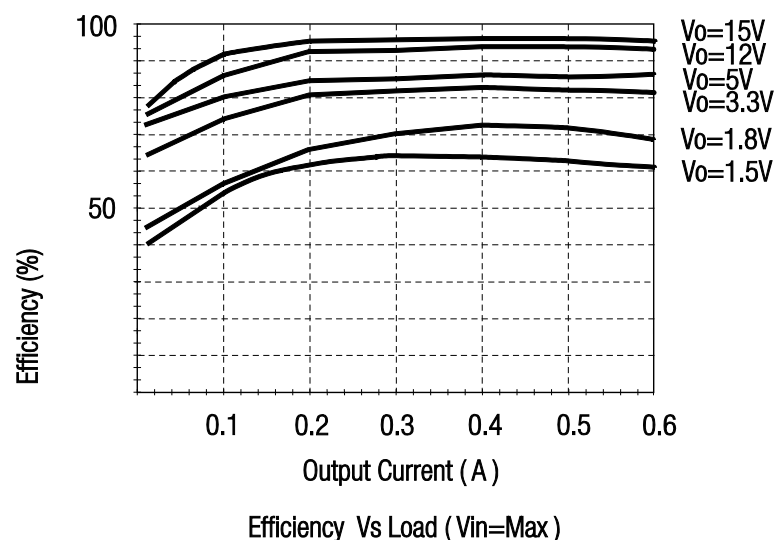


Figure 56 Plots of efficiency against output current under different regulated output voltages for the R-78XX-0.5 series from RECOM International.

Given that such high efficiencies exist for commercial switching regulators, it may be more convenient to search for one which meets the design requirements, rather than to design from scratch, depending on

the power levels in the system. As an example, consider the R-783.3-0.5 from Recom International [39]. It is small in size at 0.89 cm^3 and has a wide-input range of $4.75 - 34\text{V}$, making it suitable for regulating voltage output with significant voltage fluctuations expected from a storage capacitor.



Figure 57 Wide-input, output voltage regulator from RECOM International [39].

A linear regulator is likely to be only suitable if voltage fluctuations are minimal, otherwise the efficiency will be very low over at least part of the operating range. However, if a battery is used in the energy storage stage then a simpler and more efficient solution is probably to store the energy at a voltage which is suitable to run the load electronics directly avoiding the need for further processing and energy loss.

In summary, a wide input switching regulator is almost certainly required for a n energy harvesting system utilising capacitive storage and a battery storage system should if possible be designed so that output voltage regulation is not required.

Future Outlook

The natural progression of power electronics for energy harvesters will lead towards low-power, self-starting circuitry that would rely only on the energy scavenged from the environment. As is evident from this chapter, the condition for maximum power transfer from the harvester to the load requires continuous control of the input impedance of the interface circuitry. Future developments of the power processing stages should implement dynamic and accurate on-line tuning of the optimal damping force or adaptation of the load impedance depending on the transduction mechanism of the harvester. This has a direct consequence on the electro-mechanical coupling of the harvester and hence, the effectiveness of the harvester in converting what is deemed to be useless ambient mechanical energy into usable electrical power. Most control algorithms currently use digital signal processing in the form of microcontrollers. With the development of ultra-low power circuitry, the options available will be geared towards profoundly customised methods in the control algorithms.

The advent of highly methodical micro-fabrication techniques will provide a suitable platform allowing for the integration of energy harvesters with their power processing circuitry on a single standalone chip. In other words, energy harvesters that are compatible with MEMS technology can be easily integrated with power electronics. Recent advances in designing smaller magnetic components could significantly reduce the size of DC/DC converter circuits which forms the backbone of any adaptive impedance matching or voltage regulation circuit. Due to the high voltage and low charge characteristics of electrostatic harvesters, the power electronics design is very difficult and an unsolved problem. It is difficult to see how this will be resolved with existing semiconductor device technologies.

Conclusions

Optimisation of energy harvesters is a system level problem which involves stringent design requirements on the power processing stages. Deploying an energy harvester on its own will yield poor power densities which is why additional circuitry is needed to implement features such as an impedance match between harvester and load electronics, energy storage capabilities and output voltage regulation.

Each energy harvester is differentiated by its transduction mechanism and therefore, the equivalent source impedance model must be derived for different harvesters. By matching the source impedance to that of the load or by applying appropriate switching (as is the case for piezoelectric and electrostatic harvesters), maximum power transfer is achieved from the harvester to the load electronics under optimal conditions. This is even more crucial when energy harvesters are the potential replacements to battery-powered applications. The control overhead of the power processing stages has to be kept as low as possible to place energy harvesters in a viable position in self-powered applications. While the efficiencies of standalone, off-the-shelf power converters can reach almost 90%, this figure reduces to around 50% when the input voltage levels are within the sub-volt range. This is where the effects of

voltage drops across diodes and the power losses due to the equivalent series resistances of inductors and capacitors can negatively influence the power density of an energy harvesting system.

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