POWER MANAGEMENT FOR A BIOTELEMETRY TAG

by

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ABSTRACT

Power management for a biotelemetry tag implanted into the body of a sealion with the body temperature around 40°C is proposed in the aim for sustaining the tag’s operation for at least three years without changing power sources under the constraint of a bounded energy budget and low physical profiles of all electronic components. Various primary battery and capacitor technologies and converter topologies are studied and compared. Two types of commercially available batteries, a silver oxide battery and a lithium manganese dioxide battery, are discharged at several constant current levels and at two different temperatures to establish the Peukert equation with temperature dependence for each type of battery technology, $I^n \times t = K_0 (1 + \alpha T)$. It is concluded that a silver oxide battery is a better battery technology in that it has a more flat discharge profile and can deliver higher power and energy when they are normalized (with respect to nominal power and energy), than lithium manganese dioxide battery does.

The supercapacitor in parallel with battery cells is incorporated into the input power source to enhance the power capability required by electronic components when the battery cells alone cannot supply such high power. The supercapacitor is adopted for its highest volumetric efficiency among all capacitor technologies and relatively low leakage current, which is measured at 40°C and confirmed to be well under 0.5 $\mu$A.

A converter topology is necessary to meet different voltage requirements from all electronic components. For maximum simplicity, a converter topology using two silver oxide batteries in series producing an input voltage of 3V can involve only one boost converter in the topology based upon the unique characteristics of a silver oxide battery from test results. A commercially available compact boost converter including the battery stacks in parallel with the supercapacitor as the input power source is tested under the voltage and current requirement of a power amplifier. The result shows that the chosen boost converter can meet the input power requirement of the power amplifier.

It is concluded that under the constraint of current energy budget the proposed converter topology will operate for at most 6 years with the duration between adjacent activations for a low noise amplifier to be 46 minutes.
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CHAPTER 1
INTRODUCTION

In recent years, advances in semiconductor fabrication have made possible the proliferation and miniaturization of portable electronic devices that find use in our daily lives. Higher performance and longer service life are conflicting expectations that pose a growing challenge on the power management especially for the portable electronic devices with RF application and multiple input voltages requirement. With restricted space and energy budget, an RF portable device usually requires high power upon activation and is expected to have a long service life.

The battery is no doubt the universal power source for all kinds of portable electronic devices. A primary battery is usually the simpler and more direct choice for an input power source over a secondary or rechargeable one when the recharging means is not available or allowable. However, the battery is not ideal. The cell voltage will drop over time and the drain current is limited by internal impedance and thus may not be able to supply the high-pulse load demanded by the power amplifier. The power-enhancing component, such as the capacitor, may be adopted to meet high power requirement. Furthermore, voltage levels from the mere serial/parallel combinations of battery cells may not meet the input voltage requirements for the components in a portable electronic device. Voltage converters may be included in the power management, in order to maintain steady voltage levels demanded by those components upon activation, as the battery voltage gradually goes down.

Considering an extreme case in an implantable biotelemetry tag for tracking and monitoring
the behaviors of Stellar sealions as shown in Fig. 1, the tag is supposed to be implanted just under the scalp of a sealion, receive interrogation signals from the base station nearby, respond to interrogation signals, and is expected to last for at least 3 years without changing batteries. The board area for the tag is 30mm wide by 70mm long with a height no more than 3mm. Since a Stellar sealion can dive into the sea, all the components in the tag will have to sustain certain level of water pressure.

Fig. 1 The input power requirements and operation conditions for the three main components in an bio-telemetric tag.

There are mainly three discrete components to be assembled in the tag with their
functionalities described as follows:

1. Low noise amplifier (LNA): LNA is the ear of the tag and listens to the interrogation signals sent out from a base station nearby.

2. Power amplifier (PA): At the microcontroller’s command, the power amplifier transmits a pulse signal to the base station in response to the interrogation signals. To distinguish pulse signals from different sealions, each power amplifier of a tag has a unique transmission delay after LNA’s reception of the interrogation signals. With a transmission range of several hundred meters, the power amplifier has to be driven by high power to send out signals strong enough to reach the base station. Its duty cycle is supposed to be low, or at least lower than that of LNA, because sometimes a sealion may move out of the transmission range of a base station or dive into water and thus beyond the reach of interrogation signals.

3. Microcontroller (μC): The Microcontroller is the brain of the tag and a constant-running component. It coordinates and commands the operations of all components in the tag. When LNA receives the interrogation signals, the microcontroller acknowledges that and then commands the power amplifier and other corresponding components to transmit the respondent signals. The Microcontroller will shut down the non-active components or even disconnect them from the power sources as necessary for maximum power-savings.

The input voltage and current requirements, the active duration, and the duration between adjacent activations for the power amplifier, the low noise amplifier, and the microcontroller are
listed in Table 1. The input voltage and current requirements come from the specifications of commercially available products that are most suitable for our application at the mean time.

The active durations are fixed for all components, while the duration between adjacent activations for the power amplifier and the low noise amplifier are estimated because they are totally dependent on the total energy budget.

Table 1. Input power requirements and activation conditions for three main components in the tag

<table>
<thead>
<tr>
<th>Components</th>
<th>Voltage Level</th>
<th>Current Level</th>
<th>Active Duration/Duration between adjacent activations</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Volts</td>
<td>mA</td>
<td></td>
</tr>
<tr>
<td>Power Amplifier (PA)</td>
<td>4.75 ~ 5.25</td>
<td>1000</td>
<td>100 μS / 24 Hours (estimated)</td>
</tr>
<tr>
<td>Low Noise Amplifier (LNA)</td>
<td>2.75 ~ 3.25</td>
<td>18</td>
<td>55mS / 5 minutes (estimated)</td>
</tr>
<tr>
<td>Microcontroller (uC)</td>
<td>1.8 ~ 3.6</td>
<td>0.002 ~ 2.5</td>
<td>Constant-running</td>
</tr>
</tbody>
</table>

Ideally, LNA will remain active all the time in order not to miss any interrogation signals sent out from the base station. However, the limited energy budget cannot afford its constant-running operation. To compromise, LNA has to be turned on for a certain period of time for listening and then turned off to conserve power. The frequency of activation for LNA has to be as high as possible, or namely, the duration between adjacent activations for LNA has to be as short as possible, to increase the possibility of catching the interrogation signals, but not too high to overburden the energy budget.

The service life of the battery is strongly dependent on the temperature, and the inevitable energy losses due to the self-discharge of the battery and the leakage current of the capacitor will
be more prominent at higher temperature as in the body of a sealion.

It is the objective of this project to propose a scheme of power management for the bio-telemetric tag to ensure at least three years of service life under the constraint of tag volume and energy budget.

To achieve the objective, various technologies for batteries will be studied and compared and some of them will be tested with emphasis on the influence at elevated temperatures.

Capacitor technologies will be studied and compared with emphasis on the volumetric efficiency and the leakage current at elevated temperatures. The most suitable candidate will be tested for its leakage current at room temperature and an elevated temperature.

Converter topologies will also be studied and compared and the most suitable and simplest topology will be proposed based on the characteristics of chosen battery and capacitor technologies. A commercially available voltage converter will be chosen to be included in the proposed converter topology. This topology consisting of the chosen voltage converter and battery and capacitor technologies will be tested to ensure its capability to meet tag’s input power requirements.

Finally, the minimum duration between adjacent activations for the low noise amplifier based on the proposed converter topology and energy budget can be estimated for a specified wanted service life of the tag and the active durations listed in Table 1.
2.1 Introduction

A battery is a device that converts the chemical energy contained in its active materials directly into electric energy by means of an electrochemical oxidation-reduction reaction. In the case of a rechargeable system, the battery is recharged by a reversal of this process. While the term “battery” is often used, the basic electrochemical unit being referred to is the “cell”. A battery comprises of one or more of these cells, connected in series or parallel (or both), depending on the required output voltage and capacity.

2.2 Battery Classification

Electrochemical cells and batteries are classified as primary (non-rechargeable) or secondary (rechargeable), depending on their ability of being electrically recharged.

The primary cells or batteries are not capable of being electrically recharged, and hence, are discharged once and discarded. The primary battery is a convenient, inexpensive, lightweight source of packaged power for portable electronic and electric devices. The general advantages of primary batteries are good shelf life, high energy density at low to moderate discharge rates, little, if any, maintenance, and ease of use.

The secondary or rechargeable cells or batteries can be charged electrically, after discharge, to their original condition by passing current through them in the opposite direction to that of the
discharge current. They are storage devices for electric energy. Secondary batteries are characterized by high power density, high discharge rate, flat discharge curves, and good low-temperature performance. However, their energy densities are generally lower than those of primary batteries. Their charge retention is also poorer than that of most primary batteries.

2.3 Battery Terminologies

It is necessary to define battery-related terminologies [1] for the latter reference.

a) **Cutoff Voltage** or **End Voltage**: The prescribed voltage at which the discharge of a battery may be considered complete. Ideally, there's no charge or capacity available when the voltage of a battery drops below its cutoff or end voltage.

b) **Capacity**: The total number of Ampere-hours (Ah) or milli-Ampere-hours (mAh) that can be withdrawn from a fully charged cell or battery under specified conditions of discharge. The **Rated Capacity** is the number of Ampere-hours or milli-Ampere-hours a battery can deliver under specific conditions (rate of discharge, end voltage, temperature); usually the manufacturer's rating.

c) **C Rate**: The discharge current, in Amperes or milli-Ampere-hours, expressed as a multiple of the rated capacity in ampere-hours or milli-ampere-hours.

\[ I = M \times C_n \]

Where \( I \) = current, A or mA

\[ C = \text{numerical value of rated capacity of a battery in ampere-hours (Ah) or milli-ampere-hours (mAh)} \]
\[ n = \text{time in hours for which rated capacity is specified} \]
\[ M = \text{multiple or fraction of } C \]

For example, the 0.05C or C/20 discharge current for a battery rated at 5 Ah at the 0.2C or C/5 is 250 mA.

\[ I = M \times C_{0.2} = (0.05)(5) = 0.25 \text{ Amperes} \]

Conversely, a battery rated at 300 mAh at the 0.5C or C/2 rate, discharged at 30 mA, is discharged at the 0.1C or C/10 rate, which is calculated as follows:

\[ M = \frac{I}{C_{0.5}} = \frac{0.03}{0.3} = 0.1 \text{ or C/10} \]

d) **Depth of Discharge (DOD):** The ratio of the quantity of electricity (usually in Ah or mAh) removed from a cell or battery on discharge to its rated capacity.

e) **State of Charge (SOC):** The available capacity in a battery expressed as a percentage of rated capacity.

f) **Energy Density or Specific Energy:** The ratio of the energy output of a cell or battery to its weight (Wh/kg).

g) **Power Density or Specific Power:** The ratio of the power output of a cell or battery to its weight (W/kg).

h) **Polarization:** The change of the potential of a cell or electrode from its equilibrium value caused by the passage of an electric current.

**Activation Polarization:** That part of electrode or battery polarization arising from the
charge-transfer step of the electrode reaction.

**Concentration Polarization:** That part of electrode or battery polarization arising from concentration gradients of battery reactants and products caused by the passage of current.

**Ohmic Polarization:** That part of electrode or battery polarization arising from current flow through ohmic resistances within an electrode or battery.

i) **Overvoltage:** The potential difference between the equilibrium potential of an electrode and that of the electrode under an imposed polarization current.

### 2.4 Major Considerations for Battery Selection

A number of factors must be considered in selecting the best battery for a particular application. The characteristics of each available battery must be weighed against the equipment requirements and one selected that best fulfills these needs. The considerations that are important and influence the selection of the battery include [1]:

1. Type of Battery: Primary or secondary
2. Electrochemical System: Matching of the advantages and disadvantages and of the battery characteristics with major equipment requirements
3. Voltage: Nominal or operating voltage, maximum and minimum permissible voltages, and profile of discharge curve
4. Load Current and Profile: Constant current, constant resistance or constant power (or others); value of load current or profile, single-valued or variable load, pulsed load
5. Duty Cycle: continuous or intermittent, cycling schedule if intermittent
6. Temperature Requirements: Temperature range over which operation is required

7. Service Life: length of time operation is required

8. Physical requirements: Size, shape, weight; terminals

9. Shelf Life: Active/reserve battery system; state of charge during storage; storage time is a function of temperature, humidity and other conditions

10. Environmental Condition: Vibration, shock, spin, acceleration, etc.; atmospheric conditions (pressure, humidity, etc.)

11. Maintenance and Re-supply: Ease of battery acquisition, commercial availability, accessible distribution

In the sea-lion application, an implant-able and non-intrusive approach is required, so the primary battery type is preferred. Due to the extremely limited space of the device, the shape of the battery should be coin type or low in profile and the volumetric energy density should be as large as possible. The battery should also have a long shelf life for at least ten years and perform well at high temperature, 40 degree Celsius, and under high pressure. Table 2.4-1 ([1], [2]) lists the most commercially available batteries that could meet our application.
Table 2.4-1 Comparisons of widely used primary batteries ([1], [2])

<table>
<thead>
<tr>
<th>Type</th>
<th>Silver Oxide Battery</th>
<th>Coin type Manganese Dioxide Lithium Battery</th>
<th>Zinc Air Battery</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal Voltage</td>
<td>1.55V</td>
<td>3V</td>
<td>1.4V</td>
</tr>
<tr>
<td>End Voltage</td>
<td>1.2V</td>
<td>2V</td>
<td>1V</td>
</tr>
<tr>
<td>Positive/Negative Electrode</td>
<td>Silver Oxide/Zinc</td>
<td>Manganese Dioxide/Lithium</td>
<td>Air/Zinc</td>
</tr>
<tr>
<td>Electrolyte Solution</td>
<td>*Sodium Hydroxide</td>
<td>Organic Electrolyte</td>
<td>Potassium Hydroxide</td>
</tr>
<tr>
<td>%, capacity loss per year at 20°C</td>
<td>5</td>
<td>1-2</td>
<td>3</td>
</tr>
<tr>
<td>Energy Density</td>
<td>Volumetric</td>
<td>Gravimetric</td>
<td></td>
</tr>
<tr>
<td></td>
<td>450Wh/l</td>
<td>750Wh/l</td>
<td>1150Wh/l</td>
</tr>
<tr>
<td></td>
<td>120Wh/kg</td>
<td>260Wh/kg</td>
<td>390Wh/kg</td>
</tr>
<tr>
<td>Temp. Range</td>
<td>-10 to 60 deg.C</td>
<td>-20 to 85 deg.C</td>
<td>-10 to 60 deg.C</td>
</tr>
<tr>
<td>$SCAD$ Per Watt-hour</td>
<td>15-45</td>
<td>30-40</td>
<td>7.9</td>
</tr>
<tr>
<td>Strong Points</td>
<td>•Flat Discharge Curve •Superior Long-term Reliability •High Energy Density</td>
<td>•Stable Discharge Curve •High Energy Density</td>
<td>•For Heavy and Continuous Use •Stable Discharge Curve •Excellent Anti-leakage •High Energy Density</td>
</tr>
<tr>
<td>Limitations</td>
<td>Expensive, but cost-effective on button battery application</td>
<td>Available in small sizes</td>
<td>Drying out, flooding Limited power output</td>
</tr>
</tbody>
</table>

The Zinc/Air battery has the highest volumetric energy density (1150Wh/l) among the listed types of batteries. However, the positive electrode of this type of battery is air. An inexhaustible supply of air is necessary to be presented within the battery, which is not suitable for our
implantable application. The problem with these widely used batteries is that they are not capable of delivering high pulse current.

Table 2.4-2 ([1], [2]) lists some advanced lithium-based batteries.

Most lithium-based batteries have high voltage and superior long-term shelf life. As shown in Table 2.4-2, The Thionyl Chloride Lithium Battery \( (Li/\text{SOCl}_2) \) and Sulfur Dioxide Lithium Battery \( (Li/\text{SO}_2) \) have a safety problem due to their toxic components and therefore are not suitable for biological and medical applications. The Oxychloride Lithium Battery \( (Li/\text{SO}_2\text{Cl}_2) \) has a relatively high self-discharge rate and hence is not suitable for long-term application. Finally, The Silver Vanadium Oxide Lithium Battery \( (Li/\text{AgV}_2\text{O}_{5.5}) \) seems to be the best battery for our application due to its high-rate capability, from 1A to 2A current level, and long shelf life. Furthermore, this type of battery has been widely used in biomedical applications, such as the implantable cardioverter/defibrillator ([3]), in which high pulse current is usually needed. However, its accessibility is limited by its high cost and the discharge profile is not flat.
### Table 2.4-2 Comparison of lithium-based primary batteries ([1], [2])

<table>
<thead>
<tr>
<th>Type</th>
<th>Thionyl Chloride Lithium Battery</th>
<th>Oxychloride Lithium Battery</th>
<th>Silver Vanadium Oxide Lithium Battery</th>
<th>Sulfur Dioxide Lithium Battery</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal Voltage</td>
<td>3.6V</td>
<td>3.95V</td>
<td>3.2V</td>
<td>3V</td>
</tr>
<tr>
<td>End Voltage</td>
<td>3V</td>
<td>3.1V</td>
<td>2V</td>
<td>2V</td>
</tr>
<tr>
<td>Positive Electrode</td>
<td>Thionyl Chloride</td>
<td>Oxychloride</td>
<td>Silver Vanadium Oxide</td>
<td>Sulfur Dioxide</td>
</tr>
<tr>
<td>Electrolyte Solution</td>
<td>Non aqueous Inorganic</td>
<td>Non aqueous Inorganic</td>
<td>Non aqueous Inorganic</td>
<td>Non aqueous Inorganic</td>
</tr>
<tr>
<td>Negative Electrode</td>
<td>Lithium</td>
<td>Lithium</td>
<td>Lithium</td>
<td>Lithium</td>
</tr>
<tr>
<td>Self-discharge rate at 20 degree Celsius, percentage capacity loss per year</td>
<td>1-2</td>
<td>7 for the first year and 5 for each following year</td>
<td>2</td>
<td>2</td>
</tr>
<tr>
<td>Discharge Characteristic</td>
<td><img src="image1" alt="Graph" /></td>
<td><img src="image2" alt="Graph" /></td>
<td><img src="image3" alt="Graph" /></td>
<td><img src="image4" alt="Graph" /></td>
</tr>
<tr>
<td>Energy Density</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Volumetric</td>
<td>900 Wh/l</td>
<td>900 Wh/l</td>
<td>780 Wh/l</td>
<td>415 Wh/l</td>
</tr>
<tr>
<td>Gravimetric</td>
<td>500 Wh/kg</td>
<td>450 Wh/kg</td>
<td>270 Wh/kg</td>
<td>260 Wh/kg</td>
</tr>
<tr>
<td>Temp. Range</td>
<td>-55 to 85 deg.C</td>
<td>-55 to 85 deg.C</td>
<td>-40 to 50 deg.C</td>
<td>-55 to 70 deg.C</td>
</tr>
<tr>
<td>CAD per Watt-hour</td>
<td>520</td>
<td>-</td>
<td>550</td>
<td>-</td>
</tr>
<tr>
<td>Strong Points</td>
<td>High 3.6V Voltage</td>
<td>High 3.9 V Voltage</td>
<td>High rate capability (Amp level current)</td>
<td>High energy density</td>
</tr>
<tr>
<td></td>
<td>Flat Discharge Performance</td>
<td>Greater safety</td>
<td>Superior Long-term Reliability</td>
<td>Very flat discharge curve</td>
</tr>
<tr>
<td></td>
<td>High Energy Density</td>
<td>Higher rate capability</td>
<td>Long shelf life</td>
<td>Best low temperature, high rate performance</td>
</tr>
<tr>
<td></td>
<td>Wide Usable Temperature.</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>Superior Long-term Reliability</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Limitations</td>
<td>Voltage delay after storage</td>
<td>Cell voltage sensitive to temperature variations</td>
<td>Expensive</td>
<td>Potential safety problems</td>
</tr>
<tr>
<td>Safety Consideration</td>
<td>Higher self-discharge rate</td>
<td>Lower rate capability at low temperature</td>
<td></td>
<td>Toxic components</td>
</tr>
</tbody>
</table>
2.5 Battery Models

There are many types of batteries and many factors that affect battery performance. To predict the performance of batteries, many different mathematical models exist. None of these models are completely accurate nor do any include all necessary performance-effecting factors. Factors that affect battery performance include:

- State of charge (SOC)
- Battery storage capacity
- Rate of discharge
- Temperature
- Self-discharge
- Age/shelf life

2.5.1 Electrochemical Battery Models

The Peukert relationship states that the discharge current of a battery decreases with increasing "constant current" discharge time. Specifically (Bumby et al. [10]):

\[ I^n \times t = K \]  

Where

\[ I = \text{discharge current [amp]} \]
\[ n = \text{battery constant (n=1.35 for typical lead-acid batteries)} \]
\[ t = \text{time to discharge at current I [seconds]} \]
\[ K = \text{Constant} \]
The Peukert relationship can be written to relate the discharge current at one discharge rate to another combination of current and discharge rate:

\[ C_1 = C_2 \times \left( \frac{I_2}{I_1} \right)^{n-1} \]  \hspace{1cm} (2.5-2)

Where

- \( C \) = discharge rate

Subscripts 1 and 2 refer different discharge-rate states

From this relationship the state of charge (SOC) at a constant discharge rate is:

\[ \text{SOC} = 1 - \frac{(I \times \text{Time})}{C} \]  \hspace{1cm} (2.5-3)

For non-constant discharge rates the above equation must be modified and evaluated in small time steps:

\[ \Delta \text{SOC} = \frac{I_2 \times \text{TimeStep}}{3600} \times C_1 \times \left( \frac{I_2}{I_1} \right)^{n-1} \]  \hspace{1cm} (2.5-4)

In the above equation, it is assumed that a given combination of current and discharge rate (\( C_1 \) and \( I_1 \)) is known. Given the current at the present time step (\( I_2 \)), the corresponding discharge
rate is calculated using equation for $C_1$ and plugged into an incremental form of equation for SOC - yielding the above equation for $\Delta SOC$.

### 2.5.2 Behavioral Model

Hageman [19] developed a behavioral model for simulating the discharge behavior of common batteries using PSPICE.

The major deviation from ideal is that the usable capacity of a cell varies, depending on the discharge rate. At very low discharge rates (> 100 hours), all batteries are very efficient. At very fast discharge rates (< 10 hours), the batteries are not as efficient and usable capacity is lost. For pulsed loads with cycle times greater than 10 seconds, the cell gives more total capacity than under a constant load. The rest portion of the pulsed load allows the battery chemistry to recover some of the lost capacity. But, as the pulsed load cycle time becomes less than 1 second, the cell does not have enough time to recover and usable capacity is not increased. In these cases, the RMS value of the pulsed discharge current should be used in the simulation. The batteries are modeled in PSPICE using the following blocks (referring to the figure 2.5-1):

1. Capacitor representing the A-H capacity of the cell;
2. Discharge rate normaliser to determine the lost capacity at high discharge rates;
3. A circuit to discharge the A-H capacity of the cell;
4. Cell voltage versus state-of-charge lookup table; and
5. Cell resistance.
To start modeling a cell, several actual discharge curves should be measured at a low rate (20 to 200 hours) to get an actual voltage versus capacity curve. A single curve is then made by averaging several curves, or picking a typical curve from the data. This data is then converted into a parameterized PSpice lookup table Voltage-Controlled Voltage Source (VCVS). The actual lookup tables are composed of 30 or more pairs of data to provide finer granularity of the resulting discharge voltage curve.
To model the discharge current sense and the cell resistance, a zero-valued voltage source is added in series with the output voltage. The cell resistance is modeled as a simple resistor for NICD or Lead-Acid cells, and as a more complex variable resistance that depends on the cell’s state of charge for Alkaline cells.

To model the state-of-charge, a simple, appropriately sized capacitor is used as the charge storage element that simulates the available charge of the cell. This capacitor is sized so that it has a value of 1 Volt at 100% cell capacity and 0.5 Volts at 50% cell capacity. This capacitor is given the following value at the start of the simulation by PSpice’s “Parameterization” function: \( C_{\text{CellCapacity}} = \frac{50}{3600 \times \text{CAPACITY} \times \text{FudgeFactor}} \). The capacitor, \( C_{\text{CellCapacity}} \), is connected between nodes 50 and 0 and is given a value of the Amp-hour capacity of the cell times a conversion from hours to seconds (3,600 seconds = 1 hour) times a fudge factor (FudgeFactor). If a cell has a 10 Amp-hour capacity, \( C_{\text{CellCapacity}} \) equals 10 * 3,600 or 36,000 Farads; this is a big capacitor, but a workable value that is easy to understand.

FudgeFactor adjusts for the difference in the manufacturer’s listed Amp-hour capacity (i.e., some cutoff voltage with some capacity remaining at the cutoff) and the simulated capacity of 0 Volts output at 0% remaining capacity. To correct this, and still allow the model user to use the manufacturer’s listed capacity, a FudgeFactor value of 1.01 to 1.1 is included.

The actual usable capacity of a cell depends on the rate at which it is being discharged. Most manufacturers list the capacity at the most favorable rate—which is usually greater than 20 hours discharge. At any faster rate, the cell is less efficient and results in a nonlinear function of the discharge rate. This must be characterized as a lookup table at many discharge rates. This inefficiency is modeled as a VCVS in series with the output voltage of the battery state-of-charge.
node (the voltage on C_CellCapacity). This VCVS subtracts a given amount of capacity from the cell during discharge. The amount subtracted depends on the rate at which the cell is being discharged.

To determine the rate at which the cell is being discharged, it is convenient to normalize the discharge rate in Amps to a more conventional cell rate called the C rate. The C rate is defined as the capacity of the cell in Amp-hours when it is discharged completely in one hour. This normalization makes it easy to determine the cell inefficiency at different rates, and between different cell sizes, because it converts discharge in Amps to discharge in “C” units of the battery capacity at one hour. This conversion is done in the model by the VCVS, E_Rate, as follows:

\[
E_{\text{Rate}} \text{ RATE 0 VALUE} = \frac{\{I(V_{\text{Sense}})\}}{\text{CAPACITY}}
\]

E_Rate is the sensed discharge current in Amps divided by the Amp-hour capacity of the cell. The node, RATE, is the instantaneous rate at which the cell is being discharged (see Figure 2.5-1).

This instantaneous rate information can almost be fed directly to E_Lost_Rate to determine the actual available capacity. But, when the discharge is a low duty cycle, high value pulsed load, the cell supplies a large initial current which decays in seconds to a lower value. For pulsed loads, the cell recovers between pulses and delivers a higher proportion of its capacity than a cell under constant discharge. The delayed rate is modeled by an RC lowpass filter (R1 and C1 of Figure 146). The exact value of the RC time constant depends on the type and size of cell being simulated. E_Lost_Rate is built like the E_Cell table as follows:

\[
E_{\text{Lost\_Rate}} \text{ 50 SOC TABLE} \{ V(x) \} = (0.0,0.0) (1.5,0.5)
\]

The table entries indicate the capacity unavailable from the cell at high discharge rates. The table entry shows that at a discharge rate of 0, the cell loses 0% of its capacity (first entry). If the
discharge rate is 1.5 times the rated capacity of the cell (1.5 C), the cell loses 50% of its capacity (second entry in the table).

In Figure 2.5-1, the State-Of-Charge (SOC) node is the subtraction of the voltage on the capacitor C_CellCapacity and E_Lost_Rate: The SOC node represents the capacity in the cell for a given discharge rate during the simulation. G_Discharge discharges C_CellCapacity at the cell rate. The voltage on node 50 relates to the capacity remaining in the cell if the discharge rate is low enough to actually run the cell dry. At low discharge rates, these two nodes are the same; at high discharge rates, node SOC is at a lower potential than node 50. If, at the end of a high discharge rate the cell reverts to a low discharge, nearly the entire rated capacity can be recovered from the cell. At the high discharge rate, approximately 60% of the cell’s rated capacity can be used.

All that needs to be done now is to link the state of charge with the cell voltage to get an output. The state of charge is 1 Volt for 100%, while the cell voltage table is just the opposite. To make the cell voltage correct, the state-of-charge voltage must be inverted as shown in Figure 2.5-1.

2.5.3 Modeling Specific Factors Affecting Battery Performance

Since many conventional battery models do not handle all the factors effecting battery performance, researchers have created stand-alone models to predict the influence of these factors. Models for the factors of temperature, age and cycle history have been developed.

1) Temperature Model
The value of $K$ in Peukert's equation is known to vary with changes in temperature. Temperature corrections can be included, thus

$$I^n \times t = K_0(1 + \alpha T)$$  \hspace{1cm} (2.5-5)

Where $t$ is the discharge duration, $T$ is the temperature in °C, and $\alpha$, $K_0$, and $n$ are curve-fitting constants which are empirically determined.

Fig. 2.5-2 Behavioral battery model incorporating temperature effect proposed by Benini et al. [20] and Thottuvelil [21]

Among the various secondary phenomena that affect battery voltage, two are non-negligible: external temperature and battery internal resistance. Properly taking them into account is key to ensure model accuracy for large-capacity batteries, where the high currents delivered can cause drops and self-heating.

Temperature may impact cell behavior in many ways. The most sizable effect is due to the
offset in the output voltage caused by the heat released by the cell. This effect is particularly
evident for high discharge currents.

Hageman [19] developed a behavioral model incorporating the effect of temperature. The
same approach can also be found in Benini et al. [20] and in Thottuvelil [21]. The effect of
temperature can be modeled as a voltage loop as shown in Fig. 2.5-2. The state variable in the
thermal loop on the left ($V_{Cell,Temp}^{}$) causes an offset $V_{Temp}^{}$ in the cell output voltage. $V_{Temp}^{}$ is
obtained as the sum of the equivalent voltage ($V_e^{}$) of the environmental temperature and the
voltage source ($V_{Rise}^{}$).

$V_{Rise}^{}$ is proportional to the temperature rise due to the resistive drop $R_{th}^{}*I^2(V_{sense})^{} \theta$, where $\theta$ is the temperature rise of the cell per watt dissipated in free air. $R_{th}^{}$ lumps the effects
of both thermal resistance and thermal capacitance.

The effect of the internal resistance amounts to subtracting its resistive voltage drop from
the effective output cell voltage. The resistance $R_{int}^{}$, in the model of Fig. 2.5-2 is used to account
for such voltage drop. Other second-order phenomena influencing cell internal resistance such as
the dependence of the resistance on the cell temperature, or the state-of-charge, can be neglected.

The battery life is highly dependent of the ambient temperature. The acceleration factor $F$
follows an Arrhenius law of the form (Kervarrec and Marquet [66]):

$$F = \exp\left[\frac{E_a^{} }{R} \left( \frac{1}{T} - \frac{1}{T_{ref}^{} } \right) \right]$$

$E_a^{}$: activation energy (kilo-joules/mole)

$R$: perfect gas constant (8.3143 J/mole/K)
2) Self-discharge

The capacity loss due to the self-discharge of the battery can be substantial for a longer service time. Hence the incorporation of a self-discharge effect into the battery model is necessary for a more accurate battery life estimation. Hafen [24] has assumed that the self-discharge is a first-order reaction with a rate constant \( K \) which has an Arrhenius temperature dependence.

\[
\text{Self-discharge} = K \times \text{SOC}, \text{ where } K = \exp \left( \frac{A-B}{T_k} \right) \quad (2.5-7)
\]

2.6 Battery Life-Extending and Protection Techniques

Thomas et al. [27] proposed a hybrid energy storage system including a first energy storage device, such as a secondary or rechargeable battery, and a second energy storage device, such as an electrochemical capacitor, fuel cell, or flywheel. The second energy storage device provides intermittent energy bursts to satisfy the power requires of, for example, pulsed power communication devices. The first and second energy storage devices are coupled to a current controller to assure that pulse transients are not applied to the first energy storage device as a result of charging the second energy storage device.

Chang [28] presented a power-source-switching device automatically capable of directing power from a battery or an external power source to an all-time circuit. When the external power
source is connected, a switching circuit directs the external power source to the all-time circuit. On the other hand, when the external power source is disconnected, the switching circuit is able to direct power from the battery to the all-time circuit. Since the conventional diode connection is replaced by a switching circuitry, a voltage drop across the diode due to forward bias is avoided. Consequently, the battery can work at a low voltage level, thereby extending its working life.

All major manufacturers of primary batteries warn that the batteries may leak or explode if they are charged. The simplest means of preventing a battery from being charged from an external power source is to incorporate a blocking diode in the battery pack. The diode chosen must have a current rating in excess of the operating current of the device. It should be rated, at a minimum, at twice the operating current. The forward voltage drop should be as low as possible. Schottky diodes are commonly used because of their typical 0.5-V drop in the forward direction. Another consideration in selecting the diode is the reverse voltage rating. The peak inverse voltage (PIV) rating should be at least twice the voltage of the battery.

When multiple series stacks are paralleled within a battery pack, charging may occur when a defective or a low-capacity cell is present in one of the stacks. The remaining stacks of cells will charge the stack with the defective cell. At best, this situation will discharge the good stack, but it could result in rupture of the cells in the weak stack. To avoid this, diodes should be placed in each series string to block charging currents from stack to stack. When diode protection is used in each series stack, the diode will prevent the stack containing the defective cell from being charged. The diode should have the following characteristics ([1]):

1. Forward voltage drop should be as low as possible. Schottky type preferable.
2. Peak inverse voltage should be rated twice the voltage of the individual series stack.
3. Forward current rating of the diodes should be a minimum of

\[ I_{\text{min}} = \frac{I_{\text{op}}}{N} \times 2 \]

Where \( I_{\text{op}} \) = device operating current

\( N \) = number of parallel stacks

Fig. 2.6-1 (a) Battery circuit incorporating a blocking diode to prevent charge; (b) Series/parallel battery without diode protection; (c) Series/parallel battery with diode protection. ([1])

Batteries in "float" service are kept slightly above the gassing voltage, producing a small continuous overcharge, with gassing current more than balancing self-discharge. Series strings of batteries at the same cell current can differ slightly in gassing equilibrium, causing voltage differences between cells. Pascual and Krein [29] presented a switched capacitor network for
voltage equalization of series connected batteries. For n battery blocks, n SPDT switches are used to connect (n-1) capacitors. The switches operate together, connecting alternately to each upper and lower contact point. Each capacitor is thereby alternately connected across a battery block, and then to the adjacent block. In this way, charge is transferred from higher voltages to lower levels, and then individual battery blocks approach the same voltage over time.

Fig. 2.6-2 Cell equalization topology [1]. All the switches operate together, connecting alternately to each upper and lower contact point. Each capacitor is thereby alternately connected across a battery block, and then to the adjacent block. In this way, charge is transferred from higher voltages to lower levels, and then individual battery blocks approach the same voltage over time.
CHAPTER 3
CAPACITOR

The role of the capacitor in our application could be multifold. It can serve as an energy storage device as a part of power source to assist the battery to achieve high power/energy density. The capacitor can also be treated as a part of converter topology to be connected to the input side of a power supply to reduce electromagnetic and/or radio-frequency interference produced by the semiconductor. The capacitor can also be a part of control and logic circuitry. A wide variety of capacitors including tantalum, ceramic, film, and aluminum are used in the power supply's control circuitry. Unless used in a harsh environment, these devices are general-purpose components with low voltage and loss values.

Capacitors are an integral part of electronics technology, universally used for energy storage, wave shaping, and filtering. The basic structure consists of a dielectric material sandwiched between two electrodes or plates. The impedance of an ideal capacitor is given as $Z = X_C = 1/(2\pi fC)$ and, if graphed, would result in a straight line that gets ever closer to zero as frequency ($f$) increases. However, no circuit component is ideal, i.e., purely resistive or purely reactive. All circuit components exhibit a combination of complex impedance elements: Inductors show unwanted capacitance and hysteresis effects. Resistors display unwanted inductance characteristics. Capacitors have unwanted inductance, resistance, and dielectric absorption. Different materials and manufacturing techniques produce varying amounts of these unwanted parasitics that affect a component's performance. All components, even those constructed of the
finest materials and procedures, exhibit some of these artifacts and therefore should be modeled as complex impedances.

3.1 Considerations for the Performance of a Capacitor

There are many factors that measure the quality of a capacitor. Dissipation Factor (DF), Quality Factor (Q), and Equivalent Series Resistance (ESR) are important parameters of high performance capacitors.

**Dissipation Factor (DF)**

DF and "loss tangent" are largely equivalent terms describing capacitor dielectric losses. DF refers specifically to losses encountered at low frequencies, typically 120 Hz to 1 KHz. At high frequencies, capacitor dielectric losses are described in terms of loss tangent \(\tan \delta\). The higher the loss tangent, the greater the capacitor's equivalent series resistance (ESR) to signal power. In addition, the poorer its Quality Factor (low Q), the greater its loss (heating) and the worse its noise characteristics.

In Fig. 3.1-1, a simplified equivalent circuit model for a capacitor is shown. The capacitor can be modeled as a series combination of an inductance, L, a series resistance, Rs, and a resistor-capacitor ladder in parallel. The parallel resistor-capacitor ladder consists of the capacitance, C, and the parallel resistance, Rp. Since the parallel resistance Rp is usually the insulation resistance and is normally very high for all capacitors, the total impedance of a capacitor can be expressed as

\[
Z = \sqrt{R_s + (X_c - X_L)^2}
\]

where \(Z = \) Total impedance
Rs = Series resistance

Xc = Capacitive reactance = \frac{1}{2\pi fC}

XL = Inductive reactance = 2 \pi fL

Fig. 3.1-1 The simplified capacitor circuit model and the relationship between the loss angle, \( \delta \), and the phase angle, \( \Phi \) ([36])

When a capacitor is used as a series element in a signal path, its forward transfer coefficient is measured as a function of the dielectric phase angle, \( \Phi \). This angle is the difference in the phase between the applied sinusoidal voltage and its current component. In an ideal capacitor, \( \Phi \) equals 90°. In low-loss capacitors, it is very close to 90°.

From Fig. 3.1-1 [36], we can also derive the loss tangent as follows:
\[
\tan \delta = \left| \frac{R}{X} \right| = \frac{R_p + R_s \{1 + (\omega CR_p)^2\}}{\left(\omega CR_p\right)^2 - \omega L \{1 + (\omega CR_p)^2\}}
\]

From the above formula, \(\tan \delta\) can be derived for different frequency ranges as shown in Fig. 3.1-2 ([31]).

<table>
<thead>
<tr>
<th>MAINLY INFLUENCED BY</th>
<th>INSULATION RESISTANCE</th>
<th>POLARIZATION LOSSES</th>
<th>LEAD ELECTRODE LOSSES</th>
<th>RESONANCE</th>
<th>INDUCTION</th>
</tr>
</thead>
<tbody>
<tr>
<td>(\tan \delta = \frac{1}{\omega CR_p})</td>
<td>CRA (-\omega L)</td>
<td>CRA (-\omega L)</td>
<td>CRA (-\omega L)</td>
<td>CRA (-\omega L)</td>
<td></td>
</tr>
</tbody>
</table>

Fig. 3.1-2 Loss tangent as a function of frequency ([31])

**Quality Factor (Q)**

Quality Factor (Q) is the ratio of the energy stored to that dissipated per cycle. For a reactive component, this is defined:

\[ Q = \frac{X_c}{ESR} = \tan \Phi \]

In one aspect, Q is a figure of merit in that it defines a circuit component's ability to store
energy compared to the energy it wastes. The rate of heat conversion is generally in proportion to the power and frequency of the applied energy. Energy entering the dielectric, however, is attenuated at a rate proportional to the frequency of the electric field and the loss tangent of the material. Thus, if a capacitor stores 1000 joules of energy and dissipates only 2 joules in the process, it has a Q of 500. The energy stored in a capacitor (joules, watt-sec) = 1/2 CV².

**Equivalent Series Resistance (ESR)**

Equivalent series resistance (ESR) is responsible for the energy dissipated as heat and is directly proportional to the DF. A capacitor should be depicted as an ESR in series with an ideal capacitance (C). ESR is determined by:

\[
ESR = \frac{Xc}{Q} = \frac{Xc}{\tan \delta}, \text{ with } Q = \frac{1}{DF}.
\]

From this, we can see that "lossy" capacitors and those that present large amounts of Xc will be highly resistive to the signal power.

Circuit designs employing low Q capacitors usually produce large quantities of unwanted heat because \( \tan \delta \) and DF (or 1/Q) typically increase in a non-linear fashion with rising frequency and temperature. With some capacitors, this effect is enhanced by the naturally occurring decreased capacitance at high frequencies. High currents also produce an increase in heat, which in turn again increases the ESR and DF.

Even with substantial changes in current flow, high Q (low DF) capacitors will not exhibit the value shifts common to equivalent components exhibiting high DF, ESR, and other parasitics. Low ESR reduces the unwanted heating effects that degrade capacitors. This is an important goal in designing these components for high-current, high-performance applications, such as power supplies and high-current filter networks.
Physical Considerations

Other factors contribute additional losses: the bulk metal of leads; methods of lead attachment; capacitor plate material; and general construction. Particularly with larger capacitance values, these factors become a significant percentage of the total loss as their contribution to ESR increases.

There are three types of losses in a film capacitor: metal losses ($R_s$); leakage or insulation losses at DC or low frequencies ($R_{in}$); and dielectric losses. Leaving out extensive derivation, the following formula shows how these three losses are related to a capacitor's ESR:

$$ESR \approx R_s + \frac{1}{(2\pi f C)^2 R_{in}} + \frac{DF}{2\pi f C}$$

Impedance, Inductance, and Resonant Frequency

An ideal capacitor's reactance decreases as frequency increases, as shown by the formula:

$$X_c = 1/(2\pi f C).$$

Of course, impedance ($Z$) also varies with frequency, owing to the ESR and inductance ($L$) of the capacitor.

The point of minimum impedance (ESR) marks the frequency at which $L$ and $C$ form a series-resonant circuit, where the inductive reactance equals the capacitive reactance. Above this resonant frequency, the capacitor functions as an inductor. For many applications, the capacitor's series resonant frequency will be a circuit's useful upper frequency limit, especially where the phase angle of the capacitor is expected to maintain a 90-degree ($\tan\Phi = 0$) or near 90-degree voltage/current relationship. This is a common assumption in filter network design.
The length of the capacitor and its construction determine the capacitor's self-inductance and thus its resonant frequency. The lead length to the capacitor's external circuit load influences the in-circuit performance, usually in a quite different manner from that which was calculated based upon ideal (that is, no inductance) conditions.

**Parasitic Inductance**

The parasitic inductance of capacitors is becoming more and more important in the decoupling of today's high-speed digital systems. The relationship between the inductance and the ripple voltage induced on the DC voltage line can be seen from the simple inductance equation:

\[ V = L \frac{dI}{dt} \]

The time derivative term seen in current microcontrollers can be as high as 0.3 A/ns, and up to 10A/ns. At 0.3 A/ns, 100pH of parasitic inductance can cause a voltage spike of 30mV. While this does not sound very drastic, with the supply voltage for microcontrollers decreasing at the current rate, this can be a fairly large percentage. Another important and often overlooked reason for knowing the parasitic inductance is the calculation of the resonant frequency. This can be important for high frequency, bypass capacitors, as the resonant point will give the most signal attenuation. The resonant frequency is calculated from the simple equation:

\[ f_{\text{resonant}} = \frac{1}{2\pi \sqrt{LC}} \]
Insulation Resistance (I.R.)

Insulation Resistance is the resistance measured across the terminals of a capacitor and consists principally of the parallel resistance $R_p$ shown in the equivalent circuit. As capacitance values and hence the area of dielectric increases, the I.R. decreases and hence the product ($C \times I_R$ or $RC$) is often specified in ohm farads, or more commonly megaohm-microfarads. Leakage current is determined by dividing the rated voltage by I.R (Ohm’s Law).

3.2 Review of Capacitor Technologies

Today, capacitors come in many sizes, shapes, and varieties. I’ll only focus on the capacitor technologies that are suitable for the application of power supply. Due to our extremely limited board profile and space, surface-mounted capacitors would be the only choice. Hence, the following review for each type of capacitor technology will focus on the surface-mounted type.

a) Aluminum Electrolytic Capacitor

The construction of an aluminum electrolytic capacitor [32] requires high purity aluminum foil (for the anode and the cathode). Anode foils account for the highest costs associated with producing aluminum electrolytic capacitors. The foil is etched into tunnels that provide a surface area for the formation of aluminum oxide, which provides the capacitance. Kraft or manila separator paper is soaked in an ethyl glycol or specialty electrolyte. The paper is then rolled between the anode and cathode layers of aluminum foil. These electrolytic-grade papers are impregnated with an electrolyte (in small amounts because of high costs). Electrolytes are also produced captively by numerous capacitor manufacturers, because they can be doped with
proprietary materials that enable a company to produce a unique product line. The foil and impregnated paper layers are rolled with leads or tabs and then placed into an aluminum can. The top of the can is then sealed with a gasket assembly.

Aluminum electrolytic capacitors, while not as volumetrically efficient as other capacitors, offer the lowest cost per microfarad of the four primary dielectrics—ceramic, tantalum, aluminum and DC film. In the past, aluminum electrolytic capacitors have been known for their limited shelf life and poor low-temperature characteristics. Because of this, they are traditionally used in short shelf life electronic equipment, with emphasis upon consumer audio and video imaging products. In recent years, however, significant improvements have been realized, and both of these historical shortcomings have been improved upon, so that aluminum capacitors are now used in applications that also include computer printed circuit boards, modem cards, battery chargers for cellular phones, PDAs, air bag circuits, lighting ballasts and medical electronic devices.

b) Multilayer Ceramic Capacitor

A multilayer ceramic (MLC) capacitor [36] is a monolithic block of ceramic containing two sets of offset, interleaved planar electrodes that extend to two opposite surfaces of the ceramic dielectric. This simple structure requires a considerable amount of sophistication, both in the material and manufacturing process, to produce it in the quality and quantities required by today’s electronic equipment.
Multilayer ceramic capacitors are available in both Class 1 and Class 2 formulations. Temperature compensating formulations are Class 1 and temperature stable and general application formulations are classified as Class 2.

Class 1 – Class 1 capacitors or temperature compensating capacitors are usually made from mixtures of titanates where barium titanate is normally not a major part of the mix. They have predictable temperature coefficients and in general, do not have an aging characteristic. Thus they are the most stable capacitor available. The most popular Class 1 multilayer ceramic capacitors are C0G (NP0) temperature compensating capacitors (negative-positive 0 ppm/°C).

Class 2 – EIA Class 2 capacitors typically are based on the chemistry of barium titanate and provide a wide range of capacitance values and temperature stability. The most commonly used Class 2 dielectrics are X7R and Y5V. The X7R provides intermediate capacitance values which vary only ±15% over the temperature range of -55°C to 125°C. It finds applications where
stability over a wide temperature range is required. The Y5V provides the highest capacitance values and is used in applications where limited temperature changes are expected. The capacitance value for Y5V can vary from 22% to -82% over the -30°C to 85°C temperature range. The Z5U dielectric is between X7R and Y5V in both stability and capacitance range. All Class 2 capacitors vary in capacitance value under the influence of temperature, operating voltage (both AC and DC), and frequency.

c) Film Chip Capacitor

In a film chip capacitor using a naked and stacked construction with metallized high temperature PET (polyethylene teraphtalate), the film chip capacitor has the following advantages [34]:

- Use of high temperature dielectric films makes these capacitors suitable for IR or vapor phase reflow processes. This chip is built without specific encapsulation.
- The intrinsic elasticity of the dielectric film has excellent compatibility (of the capacitor), with all types of material for printed circuit boards.
- The self-healing property of film technology results in safe open circuit failure mode and better overall reliability.
- Excellent thermal shock resistance.
- Low dissipation factor, ESR and ESL.
- No piezoelectric effect.
- Non-polar construction.
d) Tantalum Capacitor

Tantalum capacitors [35] are manufactured from a powder of pure tantalum metal. The typical particle size is between 2 and 10 µm. The powder is compressed under high pressure around a Tantalum wire (known as the Riser Wire) to form a “pellet”. The riser wire is the anode connection to the capacitor. This is subsequently vacuum sintered at high temperature (typically 1200 - 1800°C), which produces a mechanically strong pellet and drives off any impurities within the powder. During sintering the powder becomes a sponge like structure with all the particles interconnected in a huge lattice. This structure is of high mechanical strength and density, but is also highly porous, giving it a large internal surface area. The larger the surface area, the larger the capacitance. Thus high CV/g (capacitance voltage product per gram) powders, which have a low average particle size, are used for low voltage, high capacitance parts. By choosing which powder and sinter temperature is used to produce each capacitance/voltage rating, the surface area can be controlled.

e) Niobium Oxide Capacitor

Niobium metal appears next to tantalum in the periodic table and it has similar chemical properties. Niobium ore is more abundant in its raw state and is less expensive. This has given the opportunity for tantalum capacitor manufacturers to evaluate niobium as a potential alternative to tantalum metal [35]; however there were two key barriers to niobium usage that have only been overcome recently. Firstly, the diffusion rate of oxygen from the dielectric (Nb2O5) to niobium metal is higher than tantalum, resulting in direct leakage current (DCL) instability. The second barrier was a lack of high purity niobium powders able to meet the demanding electrical and
mechanical specifications necessary for capacitor manufacture. There are two possible ways to reduce oxygen diffusion and improve DCL stability – either by doping metallic niobium powders with nitrogen or using niobium oxide powder. Niobium oxide (NbO) is a hard ceramic material characterized by high conductivity, a property usually associated with metals. Niobium oxide powder has a similar morphology to that of tantalum and niobium metals can be processed in the same way.

f) Solid Polymer Aluminum Capacitor (SPA)

The SPA capacitor [33] appears to be an offspring of a delightful marriage between the solid tantalum capacitor with the aluminum electrolytic capacitor. The SPA and its tantalum and aluminum parents create the capacitor’s dielectric as an oxide grown on a metal that connects to the positive terminal of the capacitor. However, the SPA’s positive side looks like an aluminum electrolytic and its negative side looks like a solid tantalum.

The solid tantalum capacitor starts with an oxidized sintered slug of tantalum powder with a tantalum wire. The tantalum wire connects to the positive terminal of the finished capacitor, and the tantalum oxide connects to the negative terminal through a MnO2 contact electrode and coatings of carbon and silver paint. The terminals are soldered to the capacitor element and the capacitor element is molded in resin to be the familiar tantalum chip capacitor.

The aluminum electrolytic capacitor starts with an etched and oxidized anode foil, an etched cathode foil and paper separators. Aluminum tabs connect to the foils and the foils and paper wind into a capacitor element. The element is wet with conductive electrolyte and the tabs connect to terminals in the sealed capacitor. The aluminum oxide connects to the negative
terminal through the electrolyte and cathode foil.

The SPA capacitor replaces the liquid electrolyte of the aluminum electrolytic capacitor with a solid, highly conductive polymer that becomes solid after it penetrates the pores of the etched aluminum anode foil. Then, like the solid tantalum capacitor, it connects to the negative terminal through coatings of carbon and silver paint.

The high conductance of the solid polymer sets the SPA head and shoulders above its parents. It is 1000 times more conductive than MnO2, and 10,000 times more conductive than liquid electrolyte.

![Construction Diagram of a Solid Polymer Aluminum (SPA) Chip Capacitor (3)](image)

3.3 Promising Power Sources – Supercapacitor

Energy storage devices may be broadly characterized by their energy density (energy stored per unit volume or mass) and by their power (how fast that energy can be delivered from the device).

At one end of the scale, conventional capacitors have enormous power but store only tiny
amounts of energy. At the other end, batteries can store a lot of energy but take a long time to be charged up or discharge; in other words, they have low power. Relative to these established technologies, supercapacitors offer a unique combination of high power and energy performance parameters with commercial relevance.

Batteries are 'charged' when they undergo an internal chemical reaction under a potential applied to the terminals. They deliver the absorbed energy, or 'discharge', when they reverse the chemical reaction. In contrast, when a supercapacitor is charged there is no chemical reaction. The energy is stored as a charge or concentration of electrons on the surface of a material.

This difference in principle of operation is the key to the difference in behavior and contrasting benefits of the two broad types of energy storage devices.

For many years batteries have been the preferred storage device for most applications because of their superior capability to store energy (i.e. high energy density). The amount of energy, measured in Joules, watt-hours or amp-hours, that can be stored has been sufficiently high for useful batteries to have been made and sold for all of this century. Where the application has demanded high power, the battery has been over engineered and the lifetime of the battery compromised. New battery technology such as lithium ion has been developed to increase power and energy storage. Fundamentally, however, they are energy storage devices. As such batteries will always be a poor solution where high power is required.

Supercapacitors, ultracapacitors, or electrochemical capacitors are rapidly recognized as an excellent compromise between electronic capacitors such as ceramic, tantalum and aluminum electrolytic devices and batteries. Generally, supercapacitors have energy densities several orders of magnitude higher than electronic capacitors and power densities significantly superior to
batteries as shown in Fig. 3.3-2 and 3.3-3.

3.3.1 Principles of Supercapacitor

To understand the benefits offered by supercapacitors, it is necessary to examine how an electrochemical capacitor works. The most significant difference between an electronic capacitor and an electrochemical capacitor is that the charge transfer is carried out by the electrons in the former, and by electrons and ions in the latter. The anions and cations involved in double layer supercapacitors are contained in the electrolyte which may be a liquid, (normally an aqueous or organic solution) or solid. The solid electrolyte is almost universally a conductive polymer.

Electrons are relatively fast moving and therefore transfer charge “instantly”. However, ions have to move relatively slowly from anode to cathode, and hence a finite time is needed to establish the full nominal capacitance of the device. This nominal capacitance is normally measured at 1 second.

The differences between EDLC (Electrochemical Double Layer Capacitors) and electronic capacitors are shown below ([49]):

- A capacitor basically consists of two conductive plates (electrodes), separated by a layer of dielectric material.
- These dielectric materials may be ceramic, plastic film, paper, aluminum oxide, etc.
- EDLCs do not use a discrete dielectric inter-phase separating the electrodes.
- EDLCs utilize the charge separation, which is formed across the electrode – electrolyte interface.
- The EDLC constitutes two types of charge carriers: IONIC species on the
ELECTROLYTE side and ELECTRONIC species on the ELECTRODE side.

Because highly activated carbon is used as the electrode material, each carbon particle functions as a double layer capacitor having a capacitance value of \( C_n \) (See Fig. 3.3-1, [49]).

Upon charging the capacitor, the charge has to be transferred through two resistances electronic (\( R_{e} \)) at the carbon electrode and at the carbon – current collector interface (\( R_{c} \)), and ionic (\( R_{i} \)) passing through the electrolyte. Therefore, the equivalent circuit of the EDLC is given by the above R-C combination, where \( R_{1} \), \( R_{2} \) and \( R_{n} \) are the internal resistances of the activated carbons.

![Fig. 3.3-1 Simplified equivalent circuit of an electrochemical capacitor ([49])](image)

Since the EDL capacitor is comprised of capacitors having various resistances, the charge/discharge voltage and charge/discharge time will define the apparent available capacity. Charging or discharging at a high rate may result in an apparently smaller capacitance than when done at a lower rate. This is due to the small capacitors that have large internal resistance not being fully charged or discharged. This in turn results in a large voltage drop at the start of measurement.
Fig. 3.3-2 Graphing energy density against power density, conventional batteries occupy the top-left position, and conventional aluminum-electrolytic capacitors occupy the bottom-right. Supercapacitors bridge the space between. Diagonal lines are lines of equal discharge time into a specified load ([41]).

The choice of electrolyte in an EDLC is as important as the choice of electrode material [46]. The attainable cell voltage of a supercapacitor will depend on the breakdown voltage of the electrolyte, and hence the possible energy density (which is dependent on voltage) will be limited by the electrolyte. Power density is dependent on the cell's ESR, which is strongly dependent on electrolyte conductivity. There are currently two types of electrolyte in use in EDLCs: organic and aqueous. Organic electrolytes are the most commonly used in commercial devices, due to their higher dissociation voltage. Cells using an organic electrolyte can usually achieve voltages in the range of 2 – 2.5 V. The resistivity of organic electrolytes is relatively high, however, limiting cell power. Aqueous electrolytes have a lower breakdown voltage, typically 1 V, but
have better conductivity than organic electrolytes.

![Diagram showing specific energy of capacitor types.](image)

**Fig. 3.3-3 Specific Energy of Capacitor Types (courtesy AVX, [43])**

There are two negative characteristics associated with conventional supercapacitors: high ESR in the Ohms or tens of Ohms area, and severe capacitance loss when called upon to supply very short duration current pulses. However, modern supercapacitors successfully address both of these drawbacks. This capacitance loss in the millisecond region is caused by the charge transfer (i.e. establishment of capacitance) being carried out by relatively slow moving ions in double layer capacitors. In the abovementioned "electronic" capacitors the charge transfer is performed by fast electrons, thereby creating virtually instant rated capacitance value.

Two factors are critical in determining voltage drop when a capacitor delivers a short current pulse; these are ESR and "available" capacitance, as shown in Fig. 3.3-4 ([43]).
The instant voltage drop \( V_{\text{IR}} \) is caused by and is directly proportional to the capacitor's ESR. The continuing voltage drop with time \( V_{\text{Q}} \), is a function of the available charge, i.e. capacitance. From Fig. 3.3-4, it is apparent that for very short current pulses e.g. in the millisecond region, the combination of voltage drops in a conventional supercapacitor caused by a) the high ESR and b) the lack of available capacitance, causes a total voltage drop unacceptable for most applications. Now the modern supercapacitor has an ultra-low ESR (milli-Ohm) that minimizes the instantaneous voltage drop, while the high retained capacitance reduces the charge related drop.

### 3.3.2 Supercapacitor Applications

An increasing number of portable electronic devices such as laptops and mobile phones incorporate batteries as power supplies. Many such devices draw high power, pulsed currents,
and current profiles consisting of short, high current bursts result in a reduction of battery performance. Using supercapacitors in combination with a battery is therefore an optimal solution. A supercapacitor can relieve the battery of the most severe load demands by meeting the peak power requirements, and allowing the battery to supply the average load. The reduction in pulsed current drawn from the battery results in an extended battery lifetime.

Using a supercapacitor in parallel with a battery provides a current pulse with a substantially higher voltage than that available from the battery alone. In this instance, the efficiency of the RF power amplifier is improved.

Additionally, the higher-than battery voltage supplied by the supercapacitor keeps the voltage pulse above the "cut off voltage" limit for a significantly longer time than is the case for the battery alone.

### 3.4 Comparison of Capacitors

It’s very difficult to compare all types of capacitors on the same metrics and basis. Table 3.4-1 lists the features and application circuits for the surface-mounted capacitors. Table 3.4-2 ([49]) lists the major performance characteristic for the capacitors and the supercapacitor. It should be noted that these numbers are just for reference and the real figures are dependent on the manufacturers, even for the same type of capacitor technology.
Table 3.4-1(a) Features and application circuits for various types of surface-mounted capacitors

<table>
<thead>
<tr>
<th>Type</th>
<th>Features</th>
<th>Application Circuits</th>
</tr>
</thead>
<tbody>
<tr>
<td>Aluminum electrolytic capacitor</td>
<td>This type of capacitor incorporates a metal oxide film dielectric produced from electrolysis and its anode is made of aluminum. That is the reason this type of capacitor is called aluminum electrolytic capacitor. It is possible to produce high-capacity aluminum electrolytic capacitors. Their frequency and temperature characteristics, however, are poor.</td>
<td>Power supply circuits, audio circuits, timer circuits, and backup circuits</td>
</tr>
<tr>
<td>Multi-layered ceramic capacitor</td>
<td>This type of capacitor has been developed to meet demands for high-density ceramic capacitors. Multi-layered ceramic capacitors incorporate multiple printed layers of electrode plates made of 10 to 20μm thick ceramic sheets. These capacitors are more compact and have better temperature characteristics than single-layered ceramic capacitors. Multi-layered ceramic capacitors are, however, rather expensive because their electrode plates use precious metals. With the further development of materials for electrode plates, these capacitors are expected to take the lead in the main stream of ceramic capacitors. Like a single-layered ceramic capacitor, a multi-layered ceramic capacitor is either a product of high dielectric constant construction or a product that has excellent temperature characteristics ideal for temperature compensation.</td>
<td>Circuits of general electronic equipment</td>
</tr>
<tr>
<td>Film capacitor</td>
<td>The high-frequency and temperature characteristics of film capacitors excel those of ceramic capacitors. Furthermore, high-capacity film capacitors are available, which are, however, more expensive and larger than ceramic capacitors that are the same in capacity. Polyester (Mylar), polypropylene, or polystyrene can be used for the film of this type of capacitors.</td>
<td>High-frequency circuits and analog circuits</td>
</tr>
<tr>
<td>Solid Tantalum Capacitor</td>
<td>Tantalum electrolytic capacitors are the preferred choice in applications where volumetric efficiency, stable electrical parameters, high reliability, and long service life are primary considerations. The stability and resistance to elevated temperatures of the tantalum/tantalum oxide/manganese dioxide system make solid tantalum capacitors an appropriate choice for today's surface mount assembly technology.</td>
<td>Energy storage (Power supply ripple reduction). Power filtering, bypassing. NOT suitable for timing applications.</td>
</tr>
<tr>
<td>Niobium OxideCapacitor</td>
<td>They have a similar capacitance/voltage (CV) range to current tantalum chip and demonstrate ESR characteristics comparable to conventional tantalum ratings. Their parametric stability and less expensive material cost (especially in the case of niobium oxide capacitors), make these technologies promising alternatives to low voltage tantalum and ceramic capacitors, and allow downsizing of aluminum foil capacitors. Both niobium and niobium oxide dielectrics (as well as tantalum) show no piezo effect that could degrade audio clarity if used in critical audio-video applications. The new generation of niobium and niobium oxide capacitors share the same robust casing design and industry standard sizes as current tantalum chip capacitors and are suited low ESR capacitors. However, niobium has the disadvantage of higher cost and relatively higher failure rate. The key benefits of NbO are long term stable electrical parameters, wide availability of materials, and lower cost, which should form the basis for fast designing cycles in this high growth application area.</td>
<td>The same as tantalum</td>
</tr>
</tbody>
</table>
Table 3.4-1(b) Features and application circuits for various types of surface-mounted capacitors

<table>
<thead>
<tr>
<th>Type</th>
<th>Features</th>
<th>Application Circuits</th>
</tr>
</thead>
<tbody>
<tr>
<td>Solid Polymer Aluminum Capacitor</td>
<td>Solid polymer aluminum capacitors combine the high capacitance capability of an electrolytic component with the high frequency performance of film capacitors. When the need for low impedance at high frequency is critical for one's design, one SPA chip is capable of replacing several liquid electrolyte aluminum or tantalum capacitors connected in parallel. This is due to the ultra-low e.s.r. which results in significantly lower impedance than either aluminum or tantalum capacitors at frequencies of 100 kHz and above. There is no longer a need to stack capacitors to lower the impedance at high frequency.</td>
<td>Bypassing high frequency noise, and for switching frequency filtering in DC/DC conversion.</td>
</tr>
<tr>
<td>Others</td>
<td>Mica, glass, and paper are used for dielectric elements as well as the materials described above. Mica is the best dielectric but it is expensive. Glass ensures a stable temperature coefficient in a wide range. Paper is used for high-voltage capacitors.</td>
<td>Precision equipment and high-voltage equipment</td>
</tr>
</tbody>
</table>

Table 3.4-2 Major Characteristic for Capacitors ([49])

<table>
<thead>
<tr>
<th></th>
<th>DCL</th>
<th>CV/c.c. (mFV/c.c.)</th>
<th>Reliability (hours)</th>
<th>ESR (m Ohm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Al Electrolytic</td>
<td>0.01 CV</td>
<td>6-17</td>
<td>1,000</td>
<td>Low</td>
</tr>
<tr>
<td>Ceramic</td>
<td>-</td>
<td>4-7</td>
<td>1,000</td>
<td>Low to Moderate</td>
</tr>
<tr>
<td>Film</td>
<td>-</td>
<td>8</td>
<td>1,000</td>
<td>Low</td>
</tr>
<tr>
<td>Tantalum</td>
<td>0.01 CV</td>
<td>20-30</td>
<td>1,000</td>
<td>High</td>
</tr>
<tr>
<td>Niobium</td>
<td>0.02 CV</td>
<td>20-30</td>
<td>500</td>
<td>High</td>
</tr>
<tr>
<td>Solid Polymer Al</td>
<td>0.04 CV</td>
<td>4</td>
<td>1,000</td>
<td>Low</td>
</tr>
<tr>
<td>Super Capacitor</td>
<td>2e-6 CV</td>
<td>200-400</td>
<td>25,000</td>
<td>Low</td>
</tr>
</tbody>
</table>

There are several points that should be noted in Table 3.4-2:

1. **DCL**

   It is the DC leakage current at 20°C. The leakage currents for ceramic and film capacitors are not specified in CV. They are represented by the insulation resistance, and usually in the Giga
Ω's range for both types of capacitors. Hence, the leakage currents of the ceramic and film capacitors are much smaller than other types of capacitors. The leakage current of the supercapacitor is also very small if specified by the capacitance-voltage product (CV). However, since the capacitance of the supercapacitor is usually very large compared to the conventional capacitors (22 mF is the smallest among the current search), the total leakage current will add up to 5 to 10 uA.

2. Reliability

It is the "Guaranteed Life Time" is generally used by the capacitor manufacturers to show the expected life of their capacitor products. For example, a capacitor has a Guaranteed Life Time of 20% capacitance tolerance at 105°C for 1000 hours. This means that if one charges that capacitor continuously at max rated voltage and at 105°C for 1000 hours, the capacitance will not change more than + or - 20% of it's initial value. Most of the capacitors are rated somewhere between 1000 hours to 5000 hours. Although this period of time might seem very short when compared to the length of time engineers would like their designs to function, the users need to understand that this Life Time rating is measured at the maximum rated temperature and voltage. The actual life-time of a capacitor also heavily depends on the operating temperature (ambient temperature) where the device will be used in the real world.

a) Aluminum Electrolytic Capacitor

The operating conditions directly affect the life of an aluminum electrolytic capacitor. The ambient temperature has the largest effect on life. The relationship between life and temperature follows a chemical reaction formula called Arrhenius' Law of Chemical Activity. The law, put simply, says that the life of a capacitor doubles for every 10 degree Celsius decrease
in temperature. Voltage derating also increases the life of a capacitor but to a far lesser extent, compared to temperature deratings. Internal heating caused by the applied ripple current, reduces the projected life of an aluminum electrolytic capacitor. The relationship between capacitor life and the operating conditions is expressed by the following equation [32]:

\[ L_2 = L_1 \left( \frac{V_r}{V_o} \right)^n (2)^x \]

Where \( X = \frac{T_m - T_A - \Delta T}{10} \)

\( T_A = \) Ambient temperature, \( T_m = \) Maximum operating temperature

\( \Delta T = \) Temperature rise from ripple current

\( V_r: \) Maximum rated voltage

\( V_o: \) Operating voltage

\( L_1: \) Load life rating

\( L_2: \) Projected life at operating conditions

\( n: 0 \) for axial capacitors and 1 for radial ones.

**b) Film capacitor**

The operating conditions affect the life of a film capacitor in a very similar manner to aluminum electrolytic capacitors. Voltage derating has a greater effect on the life compared to an aluminum electrolytic capacitor. The life expectancy formula for film capacitors is expressed by the following equation [34]:
c) Ceramic capacitor

The life of a ceramic capacitor at the operating conditions is expressed by the following equation [36]:

\[
L_2 = L_1 \left( \frac{V_r}{V_0} \right)^7 \cdot 2^x
\]

3.5 Comparison of supercapacitor and other types of capacitors

The supercapacitor has the advantage of higher power and energy densities, and a much higher volumetric than conventional capacitors. Table 3.4-1 lists general some comparisons of supercapacitors with conventional capacitors. Table 3.5-1 is only a general comparison. For a more specific comparison to show the excellence of the supercapacitor, a commercially available supercapacitor is compared with a tantalum capacitor. Table 3.5-2 lists the specifications of these two capacitor products.

From Table 3.5-2, we can see that the capacitance of the supercapacitor is over one hundred times higher but only seven times larger in volume than those of the tantalum capacitor. One has to connect 176 tantalum capacitors in parallel to match the capacitance of a supercapacitor. However, such a capacitor bank made of tantalum capacitors will suffer from an extremely high
leakage current, which is three thousands times higher than that of the supercapacitor with a comparable capacitance.

### Table 3.5-1 General comparison of supercapacitors with conventional capacitors ([49])

<table>
<thead>
<tr>
<th>Capacitor Technology</th>
<th>DCL (mFV/c.c.)</th>
<th>CV/c.c.</th>
<th>ESR (m Ohm)</th>
<th>Energy Density (Wh/kg)</th>
<th>Power Density (kW/kg)</th>
<th>Reliability (hours)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Al Electrolytic</td>
<td>0.01 CV</td>
<td>6-17</td>
<td>Low</td>
<td>0.04</td>
<td>4</td>
<td>1,000</td>
</tr>
<tr>
<td>Ceramic</td>
<td>-</td>
<td>4-7</td>
<td>Low to Moderate</td>
<td>0.01</td>
<td>10</td>
<td>1,000</td>
</tr>
<tr>
<td>Tantalum</td>
<td>0.01 CV</td>
<td>20-30</td>
<td>High</td>
<td>0.01</td>
<td>8</td>
<td>1,000</td>
</tr>
<tr>
<td>Super Capacitor</td>
<td>2e-6 CV</td>
<td>200-400</td>
<td>Low</td>
<td>1</td>
<td>45</td>
<td>25,000</td>
</tr>
</tbody>
</table>

### Table 3.5-2 Specifications for the supercapacitor and tantalum capacitor (courtesy cap-XX [45] and AVX [44])

<table>
<thead>
<tr>
<th>Capacitor Technology</th>
<th>Super Capacitor</th>
<th>Tantalum</th>
</tr>
</thead>
<tbody>
<tr>
<td>Manufacturer</td>
<td>Cap-XX</td>
<td>AVX</td>
</tr>
<tr>
<td>Manufacturer’s Model Number</td>
<td>GW 2 09D</td>
<td>TPME687<em>006</em>0023</td>
</tr>
<tr>
<td>Capacitance (uF)</td>
<td>120,000</td>
<td>680</td>
</tr>
<tr>
<td>Voltage (V) Nominal</td>
<td>4.5</td>
<td>6.3</td>
</tr>
<tr>
<td>ESR (mΩ) at 100 kHz</td>
<td>90</td>
<td>23</td>
</tr>
<tr>
<td>Leakage Current¹ (uA)</td>
<td>&lt; 2</td>
<td>0.01CV = 42.84</td>
</tr>
<tr>
<td>Dimensions (L x W x H mm³)</td>
<td>28.5 x 17 x 206 = 998.07</td>
<td>7.3 x 4.3 x 4.1 = 128.699</td>
</tr>
<tr>
<td>Volumetric Efficiency (mFV/c.c.)</td>
<td>541</td>
<td>33.3</td>
</tr>
<tr>
<td>Weight² (gm)</td>
<td>1.5</td>
<td>-</td>
</tr>
<tr>
<td>Maximum Pulse Current (A)</td>
<td>30</td>
<td>7</td>
</tr>
</tbody>
</table>

In addition, the total volume for such a capacitor bank would be 22 times greater than that
of the supercapacitor. For the sake of example, suppose there were a capacitor bank consisting of multiple tantalum capacitors in parallel to match the capacitance of the super capacitor. The corresponding specifications for the super capacitor and the tantalum capacitor bank of the comparable capacitance are listed in Table 3.5-3.

| Table 3.5-3 Specifications for the super capacitor and the tantalum capacitor bank of the comparable capacitance |
|---------------------------------------------------------------|---------------------------------------------------------------|
| **Super Capacitor**                                           | **Tantalum Capacitor Bank**                                   |
| Manufacturer and Model Number                                 | Cap-XX GW 2 09D                                               |
| Number of Capacitor in Parallel                              | 1                                                             |
| Capcitance (μF)                                               | 120,000                                                       |
| Voltage (V) Nominal                                          | 4.5                                                           |
| ESR (mΩ) at 100 kHz                                         | 90                                                            |
| Leakage Current (μA)                                         | <2                                                            |
| Dimensions (L x W x H mm³)                                   | 28.5 x 17 x2.06 = 998.07                                      |
| **Tantalum Capacitor Bank**                                   | AVX TPME687*006*0023                                          |
| Number of Capacitor in Parallel                              | 176                                                           |
| Capcitance (μF)                                               | 680x176 = 119,680                                             |
| Voltage (V) Nominal                                          | 6.3                                                           |
| ESR (mΩ) at 100 kHz                                         | 23/132 = 0.1307                                               |
| Leakage Current (μA)                                         | 0.01CVx 176 = 7540                                            |
| Dimensions (L x W x H mm³)                                   | 7.3 x 4.3 x 4.1 x 176 = 22651                                 |

Provided that a load current profile as shown in Fig. 3.5-1 is required for our application, the energy losses of one load cycle in either capacitor system can be expressed as:

\[
\text{Loss in one load cycle} = I_{LOAD}^2 \times ESR \times t + I_{LEAK} \times V \times T
\]  \hspace{1cm} (3-1)

The energy loss ratio of tantalum capacitor bank to supercapacitor can be expressed as:

\[
\text{Energy Loss Ratio} = \frac{\left( I_{LOAD}^2 \times ESR \times D + I_{LEAK} \times V \right)_{\text{capacitor bank}}}{\left( I_{LOAD}^2 \times ESR \times D + I_{LEAK} \times V \right)_{\text{supercapacitor}}} \hspace{1cm} (3-2)
\]
Where $D = \frac{t}{T} = \text{Duty Cycle}$, and maximum value of $D$ is one.

Fig. 3.5-1 Exemplary load current profile

![Exemplary load current profile](image)

Fig. 3.5-2 Energy loss ratio of tantalum capacitor bank to supercapacitor v.s. Duty Cycle

![Energy loss ratio graph](image)

Fig. 3.5-2 Energy loss ratio of tantalum capacitor bank to supercapacitor, versus the reciprocal of duty cycle in logarithmic scales when load current is 1.5 amps and load voltage 4.5volts

If we plug the numbers in Table 3.5-3 into (3-2) and suppose that the load current of 1.5
Amps and load voltage of 4.5V, we can get the curve of Energy Loss Ratio versus 1/D in logarithmic scales as shown in Fig. 3.5-2.

As we can see from (3-1), for greater duty cycle and thus smaller 1/D, the resistive loss due to ESR dominates the total energy loss. Therefore, the energy loss of the tantalum capacitor bank is smaller than that of the supercapacitor due to the lower ESR of the former. As the duty cycle decreases and thus 1/D increases (which is inevitably the case for most portable electronic devices and our application), the leakage loss becomes dominant in the total losses. When the duty cycle keeps decreasing to, say 1e-6, the resistive loss turns out to be negligible and the energy loss ratio comes to be stable and around the value of 2800. The transmitter in our application has to transmit 100 \( \mu \)S pulse hourly, so the duty cycle is at most 2.77e-8 and its reciprocal around 3.6e+6, which surely makes it fall into the leakage-dominant region.

It can be concluded that the supercapacitor is more suitable for applications requiring high power, low duty cycle and long service life, due to its extremely low leakage current compared with the conventional capacitors of a comparable capacitance level. Moreover, the supercapacitor also has a phenomenal advantage over its conventional counterpart in its volumetric efficiency.

3.6 Battery/Capacitor Hybrid Models

Portable electronics having wide variations in load power benefit most from adding a capacitor with a battery.

1) Hybrid Power Sources

Purdy [37] has tested the battery and the capacitor-battery power systems under pulsed
discharge conditions. It was demonstrated that the voltage drop during the pulsed current for the hybrid device was less than that for the battery, and the operating time of the hybrid device was significantly extended. The experimental results showed that the operating time and the voltage profile always improved when the capacitor was used to assist the battery. The equivalent circuit of hybrid devices for simulating voltage profile was also developed as shown in the following:

\[ I_0 = I_C + I_B \]  
\[ V_0 = V_B - I_B R_B = \left[ V_B - \frac{1}{C_{sc}} \int_0^T I_C \, dt \right] - I_C R_C \]

Voltage drop = \[ V_B - V_0 = \Delta V = \frac{I_0 R_B R_C}{R_B + R_C + \frac{T}{C_{sc}}} + \frac{I_0 R_B}{R_B + R_C + \frac{T}{C_{sc}}} \]

Wherein

\( V_0 \): load voltage
\( V_B \): battery voltage
\( I_0 \): load current
\( I_B \): battery current
\( I_C \): capacitor current
\( R_B \): battery resistance
\( R_C \): equivalent capacitor resistance
\( C_{sc} \): capacitance
The first term in (3.6-3) is the voltage drop due to the ohmic resistances, and the second term is the voltage drop due to the discharge of the capacitor. The ratio of the current delivered by the battery \(I_B\) to that by the capacitor \(I_C\) can be derived as follows:

\[
\frac{I_C}{I_B} = \frac{R_B}{R_C + \frac{T}{C_{sc}}}
\]

(3.6-4)

It can be seen that if the pulse duration is kept short and/or the capacitance is large the term \(T/C_{sc}\) is negligible and the ratio of the discharge current would be determined by the internal resistances between the battery and the capacitor.

Fig. 3.6-1 Hybrid capacitor/battery model proposed by Purdy [37], wherein (1) \(V_0\) : load voltage, (2) \(V_B\) : battery voltage, (3) \(I_o\) : load current, (4) \(I_B\) : battery current, (5) \(I_C\) : capacitor current, (6) \(R_B\) : battery resistance, (7) \(R_C\) : equivalent capacitor resistance, (8) \(C_{sc}\) : capacitance.
Maxim Application Note 671 [52] proposed a converter topology using a reservoir capacitor powers both the TDMA logic and the RF circuitry.

The capacitor supplies an average 200mA, but at 1.5A its output drop is less than 500mV after 577μs. A 1W resistor isolates the RF load from the dc-dc converter IC. While 4 x 470μF is certainly a lot of buffer capacitance in a hand-held device, the four surface-mount capacitors are far smaller and cheaper than two additional battery cells. The circuit's average power-conversion efficiency is 80%, and its quiescent supply current is only 60μA. This circuit includes a large capacitive reservoir that supplies 1.5A transient loads in a GSM cellular telephone. The average
load is only 200mA, so the 8-pin, surface-mount, boost-regulator IC requires no external MOSFET. The capacitive reservoir could be replaced by a supercapacitor to further reduce the voltage drop during transmission and the leakage current in the quiescent condition.

2) Capacitor Leakage

Under the extremely tight power budget for most portable electronic devices, the leakage current of capacitors in the circuitry cannot be ignored because the leakage current can be as high several micro-amps for the supercapacitors or at high temperature.

Franklin [47] [48] assessed the relative importance of factors leading to high leakage from measurements of leakage over a range of test conditions.

Even though it is normally quoted as such, leakage current is not one single value; it varies markedly with time, voltage, and temperature, and also has a distinct history dependence.

Effect of Time

The plot of leakage current (I) against time (t) for the first 5 minutes of electrification can often be approximated as the sum of three components (a) the charging current (log I = A – Bt); (b) a current which is inversely proportional to time (I*t = constant); and (c) a current which changes only slowly with time. The charging current drops logarithmically with time and is usually negligible after a few seconds except for the highest capacitance values. In any case, it can be calculated from the capacitance and the charging resistance and so allowed for in the analysis. Within any batch the I*t term tends to be similar from one capacitor to another. The fault current, which is what remains after deducting the charging current and the I*t current for the total leakage current, is the value which varies most between capacitors.
When the time scale is extended beyond five minutes, it is seen that this residue is not actually constant; normally it slowly drops with time although, in some instances, it can increase. As a first approximation the leakage can be split into the two components for further analysis as shown in Fig. 3.6-3.

![Diagram showing leakage current versus time for a 47 uF 35V tantalum capacitor. The leakage current consists of two parts, the fault current which is independent of time, and the I*t current which fades out with time.](image)

Effect of Voltage

From what is known already about dielectric absorption, it could be assumed that the discharge current is directly proportional to voltage. This has in fact been found to be true (Fig. 3.6-4).

Generally, the fault current is more sensitive to voltage, increasing typically 100 to 1000 fold for a 10-fold voltage increase. This difference in voltage sensitivity can result in situations
where the leakage at rated voltage is mainly due to the fault current while that at 0.1 x rated voltage is mainly dielectric absorption. Unless the two components are separated, the true effect of voltage cannot be assessed. Another aspect of voltage is the relationship to the rated voltage of the capacitor. In the measurements on our own product, the discharge current has been found to be lower in terms of nA/µFV the higher the rated voltage.

![Discharge Current vs. Normalized Current](image)

**Fig. 3.6-4 Effect of voltage on the leakage current for a 150 uF, 16V tantalum capacitor ([47])**

**Effect of Temperature**

The effect of temperature is complicated by significant deviations from the \( I^t = \text{constant} \) relationship at high temperature as shown in Fig. 3.6-5.

To summarize the Causes of High Leakage Current for the tantalum capacitor so far:

- A single valued leakage current is of no value for fault analysis;
- The leakage current increases with temperature;
- The leakage current increases with the applied voltage;
In order to reduce the capacitor leakage, the capacitor has to be operated under lower temperature and voltage. For example, a typical tantalum capacitor has its leakage current doubled at 40°C compared with that at 20°C and quintupled at rated voltage compared with that at half of the rated voltage.

Conway [30] addresses the issue of the temperature effects on self-discharge for supercapacitors. Self-discharge is a rate process so it is affected by temperature according to well-known kinetic effects associated with an activation energy, $\Delta E$. Generally self-discharge current, $I$, will be affected by temperature according to an Arrhenius relation of the form

$$I(T) = Ae^{-\frac{\Delta E}{RT}}$$

(3.6-5)
With an activation energy of, e.g., 40 kJ mol\(^{-1}\) (a typical value), the ratio of self-discharge currents at 37°C (310K) compared with 20°C(293K) is easily calculated from equation (5.2-5), i.e.,

\[
\frac{i(37°C)}{i(20°C)} \equiv \frac{i(310K)}{i(293K)} = e^{\left(\frac{-40,000}{8.31} \left[\frac{1}{310} - \frac{1}{293}\right]\right)} = 2.46
\] (3.6-6)

For example, the leakage current for cap-XX GW214D supercapacitors in Chapter 3 is no greater than 2 \(\mu\)A at 20°C according to the manufacturer’s data. We will expect a higher leakage current of 5 \(\mu\)A when the supercapacitor operates at 37°C, which is the case for our application.

Note also that these temperature effects may not be identical at each of the two electrodes of an electrochemical capacitor cell, since different self-discharge mechanisms may operate at the positively and negatively charged electrode interfaces. This is probably more usually the case for battery self-discharge, where the anode and cathode reactions are normally quite different chemically.
4.1 Linear Regulator

The linear regulator is the original form of the regulating power supply. It relies upon the variable conductivity of an active electronic device to drop voltage from an input voltage to a regulated output voltage. In accomplishing this, the linear regulator wastes a lot of power in the form of heat, and therefore gets hot. It is, however, a very electrically “quiet” power supply. Linear regulators are step-down regulators only; that is, the input voltage source must be higher than the desired output voltage. There are two types of linear regulators: the shunt regulator and the series-pass regulator. The shunt regulator is a voltage regulator that is placed in parallel with the load. An unregulated current source is connected to a higher voltage source, the shunt regulator draws output current to maintain a constant voltage across the load given a variable input voltage and load current. A common example of this is a Zener diode regulator. The series-pass linear regulator is more efficient than the shunt regulator and uses an active semiconductor as the series-pass unit between the input source and the load. The series-pas unit operates in the linear mode, which means that the unit is not designed to operate in the full on or off mode but instead operates in a degree of “partially on.” The negative feedback loop determines the degree of conductivity the pass unit should assume to maintain the output voltage.

The heart of the negative feedback loop is a high gain operational amplifier called a voltage error amplifier. Its purpose is to continuously compare the difference between a very stable
voltage reference and the output voltage. A stable voltage reference is placed on the non-inverting input of the voltage error amplifier. The gain of the voltage error amplifier produces a voltage that represents the greatly amplified difference between the reference and the output voltage (error voltage). The error voltage directly controls the conductivity of the pass unit thus maintaining the rated output voltage. If the load increases, the output voltage will fall. This will then increase the amplifier's output, thus providing more current to the load.

Fig. 4.1-1 The topology for the basic linear regulator ([51])

4.2 Switching Regulator

The switching power supplies are more efficient and are smaller in size than linear regulators of similar ratings. They are, however, more difficult to design and radiate more electromagnetic interference (EMI). Unlike linear regulators which operate the power transistor in the linear mode, the switching power supply operates the power transistor in both the saturated and cutoff states. In these states, the voltage-ampere product across the power transistor is always kept low (saturated, low-\(V\)/high-\(I\); and cutoff, high-\(V\)/no-\(I\) ). This EI product within the power
device is the loss within all the power semiconductors. The more efficient operation of the switching power supply is done by "chopping" the direct current (dc) input voltage into pulses whose amplitude is the magnitude of the input voltage and whose duty cycle is controlled by a switching regulator controller.

There are two major operational types of switching power supplies: the forward-mode converter and the boost-mode converter. Their topologies are shown in Fig. 4.2-1 (a) and (b).

a) The Forward-mode Converter

The operation of the forward-mode converter can be seen as analogous to a mechanical flywheel and a one-piston engine. The L-C filter, like the flywheel, stores the energy between the
power pulses of the driver. The input to the L-C filter (choke input filter) is the chopped input voltage. The L-C filter volt-time averages this duty-cycle modulated input voltage waveform.

The L-C filtering function can be approximated by

\[ V_{\text{out}} \approx V_{\text{in}} \cdot \text{dutycycle} \]  \hspace{1cm} (4.2-1)

The output voltage is maintained by the controller by varying the duty cycle. This buck converter is also known as a step-down converter, since its output must be less than the input voltage.

The operation of the buck regulator can be seen by breaking its operation into two periods (refer to Fig. 4.2-2). When the switch is turned on, the input voltage is presented to the input of the L-C filter. The inductor current ramps linearly upward and is described as

\[ i_{L(on)} = \frac{(V_{\text{in}} - V_{\text{out}}) \cdot t_{\text{on}}}{L_o} + i_{\text{initial}} \]  \hspace{1cm} (4.2-2)

The energy stored within the inductor during this period is

\[ E_{\text{stored}} = \left( \frac{1}{2} \right) L_o \left( i_{pk} - i_{\text{min}} \right)^2 \]  \hspace{1cm} (4.2-3)

This input energy is stored by the flux contained within the core material of the inductor.

When the power switch is turned off, the input voltage to the inductor wants to fly below ground and the diode (D), called a catch diode, becomes forward biased. This continues to conduct the current that was formerly flowing through the power switch and some of the stored energy is discharged to the load. This forms a local current loop that includes the diode, inductor,
and the load.

![Diagram of voltage and current waveforms for a forward-mode converter (buck converter, [51])](image)

The current through the inductor is described during this period by

\[
i_{L(\text{off})} = i_p - \frac{V_{\text{out}} \cdot t_{\text{off}}}{L_o}
\]

The current waveform, this time, is a negative linear ramp whose slope is \(-V_{\text{out}} / L\). When the power switch once again turns on, the diode snaps off and the current flows through the input power source and the power switch. The inductor’s current (imin) just prior to the switch being turned on, becomes the initial current the power switch must then initially pass.

b) The Boost-mode Converter

The boost-mode converter has the same parts as the forward-mode converter, but they have
been re-arranged. When the power switch is turned on, a current loop is created that only includes the inductor, the power switch, and the input voltage source. The diode is reversed-biased during this period. The inductor’s current waveform is also a positive ramp and is described by

$$i_{L(\text{off})} = i_{pk} - \frac{V_{\text{out}} \cdot t_{\text{off}}}{L_o}$$

(4.2-5)

Energy is stored in the flux within the inductor’s core material. When the power switch is turned off, the inductor’s voltage, ‘flies back’ above the input voltage. The diode becomes forward-biased when the inductor’s voltage exceeds the output voltage. The inductor’s voltage is then clamped at the value of the output voltage. This voltage level is referred to as the flyback voltage and is the value of the output voltage plus one diode forward voltage drop. The inductor current during the power switch’s off-period is described by

Fig. 4.2-3 Waveforms for a continuous-mode boost converter ([51])
When the core’s flux is completely emptied prior to the next cycle, it is referred as the **discontinuous-mode** of operation as shown in Fig. 4.2-4. When the core does not completely empty itself, a residual amount of energy remains in the core. This is called the continuous mode of operation and can be seen in Fig. 4.2-3. The majority of boost-mode converters operate in the discontinuous mode since there are some intrinsic instability problems when operating in the **continuous mode**.

\[ i_{L(\text{off})} = i_{pk(\text{on})} - \frac{(V_{\text{out}} - V_{\text{in}}) \cdot t_{\text{off}}}{L} \]  \hspace{1cm} (4.2-6)

The energy stored within the inductor of a discontinuous-mode boost converter is given by
The energy delivered per second, or watts, must be sufficient to meet the continuous power demands of the load. This means that the energy stored during ON time of the power switch must have a high enough \( I_{pk} \) to satisfy equation 4.2-7:

\[
E_{\text{stored}} = \left( \frac{1}{2} \right) L_o \left( I_{pk} \right)^2
\]  

Table 4.2.1 lists the comparison for various topologies of the switching regulators.

Discontinuous conduction lets the inductor current decay to zero during each off period, which causes a transfer of all the stored energy to the output filter during each switching cycle. In the continuous-conduction mode, the inductor current includes a DC component proportional to the load. Operating in the continuous-conduction mode lowers the ratio of peak inductor current to DC-load current, thereby lowering the peak-to-peak ripple current and reducing the core loss.

The parameters affecting the choice of a switching-regulator topology include the peak currents for the load and the inductor, the voltage level on the power transistors, and the necessity for magnetic and capacitive energy storage.

In battery-powered converters, the peak inductor current is important, because it affects the battery life and parasitic losses. The peak currents depend partly on the average load current,
which varies with the regulator topology. Peak inductor current also depends on the control circuit and whether the inductor current is continuous. The voltage stress on the switching transistor is usually not an issue in battery-powered converters. The 20V and 50V breakdown voltage ratings for standard logic-level MOSFETs are adequate for the low input and output voltages found in battery-powered systems.

Table 4.2-1 Comparison of the switching regulator topologies ([51])

<table>
<thead>
<tr>
<th>Topology</th>
<th>Power Range</th>
<th>Vin(dc)</th>
<th>In/Out</th>
<th>Efficiency</th>
<th>Parts</th>
</tr>
</thead>
<tbody>
<tr>
<td>Buck</td>
<td>0-1000</td>
<td>5-40</td>
<td>No</td>
<td>78</td>
<td>1.0</td>
</tr>
<tr>
<td>Boost</td>
<td>0-150</td>
<td>5-40</td>
<td>No</td>
<td>80</td>
<td>1.0</td>
</tr>
<tr>
<td>Buck-Boost</td>
<td>0-150</td>
<td>5-40</td>
<td>No</td>
<td>80</td>
<td>1.0</td>
</tr>
<tr>
<td>IT forward</td>
<td>0-150</td>
<td>5-500</td>
<td>Yes</td>
<td>78</td>
<td>1.4</td>
</tr>
<tr>
<td>Flyback</td>
<td>0-150</td>
<td>5-500</td>
<td>Yes</td>
<td>80</td>
<td>1.2</td>
</tr>
<tr>
<td>Push-pull</td>
<td>100-1000</td>
<td>50-1000</td>
<td>Yes</td>
<td>75</td>
<td>2.0</td>
</tr>
</tbody>
</table>

Dissipation losses occur in the parasitic resistive elements of the regulator circuit. These losses include the series resistance of the battery, the equivalent series resistance (ESR) of the filter capacitors, the on-resistance of the switching element, and the resistances in the conductors, connectors, and wiring. Dissipation losses are proportional to the square of the peak current, so
reducing the peak current can greatly minimize these losses. In addition, internal heating degrades a battery's chemistry; thus, excessive peak currents can shorten a battery's life.

The best choice for most battery-powered applications is the buck regulator, provided that one can afford the several cells needed to generate a battery voltage higher than the output voltage.

Inductor current flows to the load during both phases of the switching cycle, so the average output current equals the average inductor current. In theory, the best efficiency occurs when the input voltage is low, which implies less battery cells in series. Assuming the switch's on-state voltage drop is much smaller than the input voltage, a low input voltage reduces the AC switching losses and the RMS input current.

The boost, or step-up, topologies generate an output voltage that is greater than the input voltage. The boost topologies suit systems with a limited number of battery cells. Because the source voltage and the inductor are in series, the average inductor current equals the DC input current given by:

$$I = \frac{P_{in}}{V_{in}}.$$

Sometimes called the buck-boost circuit, the inverter topology generates an output voltage that has opposite polarity from the input voltage. Inverting and fly-back regulators are electrically equivalent when considering peak currents and voltage stress. These topologies are most suitable to applications that require negative or galvanically isolated outputs. In general, however, the high peak currents make inverting and flyback topologies the least attractive of the simple regulators.
4.3 Inductor-less Converter

Despite the advances made in inductor-based switching regulators, most designers would regard the ideal converter circuit as one that has no inductor. The capacitor-based alternatives (charge-pump converters) were hampered in the past by their lack of regulation and limited output current. Though still low compared to that of switching regulators, their output current is now adequate for many designs. In some cases, the charge-pump advantages are compelling - low cost, small size, and reduced EMI.

Charge-pump voltage converters [54] [55] use ceramic or electrolytic capacitors to store and transfer energy. Although capacitors are more common and much cheaper than the coils used in other types of DC/DC converters, capacitors can't change their voltage level abruptly. A changing capacitor voltage always follows the exponential function, which imposes limitations that inductive voltage converters can avoid. On the other hand, inductive voltage converters are more expensive.

Capacitive voltage conversion is achieved by switching a capacitor periodically. Passive diodes can perform this switching function in the simplest of cases, provided an alternating voltage is available. Otherwise, DC voltage levels require the use of active switches, which first charge the capacitor by connecting it across a voltage source and then connect it to the output in a way that produces a different voltage level. The charge pump integrates switches and the oscillator so that the switches S1, S3 and S2, S4 work in alternation (Fig. 4.3-1). The configuration shown inverts the input voltage. With a slight change in the external connections, it can double or divide the input voltage as well.
Fig. 4.3-1 The configuration for a charge-pump inverting converter [54]. Closing S1 and S3 charges the flying capacitor (C1) to V+ in the first half cycle. In the second half, S1 and S3 open and S2, S4 close. This action connects the positive terminal of C1 to ground and connects the negative terminal to VOUT. C1 is then in parallel with the reservoir capacitor C2. If the voltage across C2 is smaller than that across C1, charge flows from C1 to C2 until the voltage across C2 reaches -(V+).

Closing S1 and S3 charges the flying capacitor (C1) to V+ in the first half cycle. In the second half, S1 and S3 open and S2, S4 close. This action connects the positive terminal of C1 to ground and connects the negative terminal to VOUT. C1 is then in parallel with the reservoir capacitor C2. If the voltage across C2 is smaller than that across C1, charge flows from C1 to C2 until the voltage across C2 reaches -(V+).

The building block and the equivalent circuit for a charge pump converter are shown in Fig. 4.3-2 (a) and (b), respectively. With the switch in position A of Fig. 4.3-2(a), capacitor C1 will charge to voltage V1. The total charge stored on C1 is q1 = C1*V1. The switch is then flipped to position B discharging C1 to voltage V2. The charge remaining on C1 is q2 = C1*V2.
transferred to the output V2 is, therefore, the difference between q1 and q2, so \( q = q_1 - q_2 = C_1 (V_1 - V_2) \).

As the switch is toggled between A and B at a frequency f, the charge transfer per unit time or current is: \( I = f (q) = f (C_1)(V_1 - V_2) \). Therefore, \( I = (V_1 - V_2)/(1/ fC_1) = (V_1 - V_2)/(R_{eq}) \), where \( R_{eq} = 1/fC_1 \). The switched capacitor may, therefore, be replaced by an equivalent resistance whose value is dependent on both the capacitor size and the switching frequency. This explains why lower capacitor values may be used with higher switching frequencies. It should be remembered that as the switching frequency is increased the power consumption will increase due to some charge being lost at each switching cycle. As a result, at high frequencies, the power
efficiency starts decreasing. Other losses include the resistance of the internal switches and the equivalent series resistance (ESR) of the charge storage capacitors.

The drawbacks of the charge pump converter are their unregulated output voltage and limited current capability, which barely exceeds 100mA. A linear regulator built within the charge pump can provide a regulated output, but the efficiency drops accordingly to no more than 90% for most charge pump converters.

4.4 Comparison of Converter Topologies

The parameters most significant in power-supply design are cost, efficiency (battery life), output ripple and noise, and quiescent current. Table 4.3-1 [52] illustrates the trade-off among these parameters for five power-supply architectures and five combinations of VIN/VOUT range. In drawing attention to the strengths and weaknesses of these architectures, the following discussion also points to some surprising results from the table.

Low-Dropout (LDO) Linear Regulators

The LDO's lowest cost, lowest noise, and lowest quiescent current make it a solid choice for many applications. Its external components are minimal: usually a bypass capacitor or two. The newest LDOs offer dramatically improved performance, though of course not all the following are available in the same device: 30μVrms output noise, 60dB PSRR, 6μA quiescent current, and 100mV dropout. Efficiency, though poor when VIN is much larger than VOUT, becomes very high when VIN approaches VOUT. In that case, the LDO benefits are almost impossible to surpass. In fact, many circuits converting Li-ion battery voltage to 3V use an LDO, despite having to discard 10% or more of the battery's capacity at the end of discharge. Despite this
compromise, LDO circuits for this application offer the longest battery life among low-noise architectures.

Table 4.3-1 Power-supply performance vs. architecture vs. VIN/VOUT Range ([52])

<table>
<thead>
<tr>
<th>VIN &gt;&gt; VOUT</th>
<th>Low Cost</th>
<th>High Efficiency</th>
<th>Low Noise</th>
<th>Low Current</th>
</tr>
</thead>
<tbody>
<tr>
<td>LDO Linear</td>
<td>A</td>
<td>D</td>
<td>A</td>
<td>A</td>
</tr>
<tr>
<td>Charge Pump Reg.</td>
<td>B</td>
<td>B</td>
<td>C</td>
<td>C</td>
</tr>
<tr>
<td>Charge Pump + LDO</td>
<td>C</td>
<td>B</td>
<td>A</td>
<td>D</td>
</tr>
<tr>
<td>DC-DC Buck</td>
<td>C</td>
<td>A</td>
<td>C</td>
<td>B</td>
</tr>
<tr>
<td>Buck + LDO</td>
<td>D</td>
<td>B</td>
<td>A</td>
<td>C</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>VINmin = VOUT</th>
<th>Low Cost</th>
<th>High Efficiency</th>
<th>Low Noise</th>
<th>Low Current</th>
</tr>
</thead>
<tbody>
<tr>
<td>LDO Linear</td>
<td>A</td>
<td>B</td>
<td>A</td>
<td>A</td>
</tr>
<tr>
<td>DC-DC Buck</td>
<td>B</td>
<td>C</td>
<td>A</td>
<td>B</td>
</tr>
<tr>
<td>Buck + LDO</td>
<td>C</td>
<td>D</td>
<td>A</td>
<td>C</td>
</tr>
<tr>
<td>Boost + LDO</td>
<td>D</td>
<td>B</td>
<td>A</td>
<td>C</td>
</tr>
<tr>
<td>Buck/Boost</td>
<td>D</td>
<td>D</td>
<td>D</td>
<td>D</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>VINmin &lt; VOUT &lt; VINmax</th>
<th>Low Cost</th>
<th>High Efficiency</th>
<th>Low Noise</th>
<th>Low Current</th>
</tr>
</thead>
<tbody>
<tr>
<td>Charge Pump Reg.</td>
<td>B</td>
<td>C</td>
<td>C</td>
<td>C</td>
</tr>
<tr>
<td>Charge Pump + LDO</td>
<td>C</td>
<td>D</td>
<td>A</td>
<td>D</td>
</tr>
<tr>
<td>Boost + LDO</td>
<td>D</td>
<td>A</td>
<td>B</td>
<td>C</td>
</tr>
<tr>
<td>Buck/Boost</td>
<td>D</td>
<td>B</td>
<td>D</td>
<td>C</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>VINmax = VOUT</th>
<th>Low Cost</th>
<th>High Efficiency</th>
<th>Low Noise</th>
<th>Low Current</th>
</tr>
</thead>
<tbody>
<tr>
<td>Charge Pump Reg.</td>
<td>B</td>
<td>C</td>
<td>C</td>
<td>C</td>
</tr>
<tr>
<td>Charge Pump + LDO</td>
<td>C</td>
<td>C</td>
<td>A</td>
<td>D</td>
</tr>
<tr>
<td>DC-DC Boost</td>
<td>C</td>
<td>A</td>
<td>D</td>
<td>B</td>
</tr>
<tr>
<td>Boost + LDO</td>
<td>D</td>
<td>B</td>
<td>B</td>
<td>C</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>VIN &lt;&lt; VOUT</th>
<th>Low Cost</th>
<th>High Efficiency</th>
<th>Low Noise</th>
<th>Low Current</th>
</tr>
</thead>
<tbody>
<tr>
<td>Charge Pump Reg.</td>
<td>B</td>
<td>C</td>
<td>C</td>
<td>C</td>
</tr>
<tr>
<td>Charge Pump + LDO</td>
<td>C</td>
<td>C</td>
<td>A</td>
<td>D</td>
</tr>
<tr>
<td>DC-DC Boost</td>
<td>C</td>
<td>A</td>
<td>D</td>
<td>B</td>
</tr>
<tr>
<td>Boost + LDO</td>
<td>D</td>
<td>B</td>
<td>B</td>
<td>C</td>
</tr>
</tbody>
</table>

A = excellent, B = good, C = average, D = poor
Charge Pumps

The basic charge pump offers low cost, requires only a few external capacitors, and is usually approximately 95% efficient. Constant switching action, however, produces output noise and high quiescent current. As another issue, charge pump outputs produce only exact multiples of the input voltage. With, for example, four internal switches and one external flying capacitor, these multiples are limited to +2x, +1/2x, and 1x. Doubling the circuitry makes other multiples available, at the cost of reduced output power or of greater expense and quiescent current. Basic charge pumps seldom connect directly to the battery. Instead, they usually generate secondary voltages from existing regulators.

Charge Pump plus LDO

The charge pump plus LDO architecture avoids the problem of exact voltage multiplication. It also reduces output noise, but at the expense of efficiency. This efficiency loss can be small or large, depending on the relative magnitudes of input and output voltage. As an example, the efficiency for converting a two-cell NiMH battery to 3V output is calculated as follows:

Charge-pump efficiency is not a factor in this expression, because any such efficiency less than 100% causes the charge-pump output to drop, which lowers the input voltage to the LDO and thereby improves the LDO's efficiency.

Charge Pump Regulators

By employing pulse-frequency modulation (PFM) or pulse-width modulation (PWM), the newer charge pump regulators dispense with the need for an LDO. Compared with the charge pump/LDO approach, a regulated charge pump costs less and offers lower quiescent current in the PFM mode, but it has the same efficiency and greater output noise. Some implementations
improve efficiency by changing the multiplication factor as needed. As an example, a conversion from a two-cell alkaline battery to 5V uses +2x multiplication when the batteries are fresh and switches automatically to +3x when the battery voltage falls below 2.5V. In a buck/boost application, another charge pump might start with +1x for buck and switch to +2x for boost. Regulated charge pumps of this sophistication are still relatively rare in the semiconductor industry.

**DC DC Converters**

Available in buck, boost, buck/boost, and inverting topologies, DC DC converters offer high efficiency, high output current, and medium-low quiescent current. On the other hand, they produce output ripple and switching noise. They also are more expensive, thanks to their more complicated control schemes and the need for an external inductor. In recent years, the push toward sub-micron chip fabrication has reduced the cost penalties in several ways. First, the lower on-resistance in MOSFETs has, in many applications, eliminated the need for external FETs by enabling higher output power. It's now possible, for example, for a boost converter with 3.6V input and on-chip NFET to produce an output of 2A at 5V. Second, the small die size intended for low-to-medium-power applications allows the use of small and inexpensive packaging. Third, faster switching frequencies (up to 1MHz) have reduced the cost and physical size of external capacitors and inductors. Finally, better control schemes have added valuable features such as soft-start capability, current limiting, and selectable PWM or PFM operation.

**DC DC Buck Converter**

In nearly all applications for which VIN is greater than VOUT, the DC DC buck converter is more efficient than an LDO. This is especially true when VIN is much greater than VOUT, as,
for instance, in converting the output of a single Li-ion cell to 1.8V. The DC DC buck converter exhibits some output ripple and switching noise, but these artifacts are not as severe as in other DC DC topologies. One notable advance in control schemes is the implementation of duty cycles up to 100%, enabling the circuit to achieve low-dropout performance.

**DC DC Buck Converter with LDO**

Combining the DC DC buck converter with an LDO is useful in applications for which high efficiency and low noise are priorities. This arrangement, however, applies only when VIN is substantially larger than VOUT. If the minimum VIN approaches VOUT, the LDO alone should provide similar efficiency and lower dropout, usually resulting in the same or better battery life at a much lower cost.

**DC DC Boost Converter**

The most important feature of a DC DC boost converter is that an LDO cannot perform the same function. The closest competition is the regulated charge pump, which has lower efficiency and lower output power. On the other hand, boost converters have notoriously high output ripple and switching noise. They also require better control schemes, to eliminate oscillation in the output and to reduce efficiency loss due to parasitic resistance in the MOSFET switch and external components.

**DC DC Boost Converter plus LDO**

Combining a DC DC boost converter with an LDO has two advantages: It implements a low-noise boost function (at a slight penalty in efficiency versus the noisy booster without an LDO), and it performs the buck/boost function with surprisingly high efficiency. A typical buck/boost application converts the output of one Li-ion cell to 4.3V. Efficiency is very high,
because the battery spends most of its life near 3.6V, allowing the booster to idle and providing the LDO with a near-ideal input voltage. This system also delivers higher efficiency with smaller external components than the traditional SEPIC converter. Because of the favorable characteristics of this arrangement, several single-chip implementations are available for the DC-DC boost converter plus LDO architecture.

Table 4.3-2 ([58]) shows the most common regulator topologies beginning with the simplest (top of table) (which are generally also the most desirable in terms of efficiency, cost, and size) and progressing to more specialized types (bottom). The table also lists the pros and cons of each topology, allowing designers to scroll through and find the topology that best meets their needs.

In our application, the battery can directly drive the microcontroller due to its low current requirement and wide input voltage range. With the help of capacitor it is still possible to drive the low noise amplifier without a voltage converter. Now, our concern boils down to the power amplifier. The mere serial or parallel combinations of battery cells seem not be able to generate a 5V output required to drive the power amplifier. Hence, at least a voltage converter for the power amplifier will be involved in our converter topology.

The converter suitable for our application should have the following features:

1. High efficiency to make the best use of available energy.
2. Accepting low input voltage (< 2V) so that it can still operate normally even when the input battery voltage drops below the cutoff voltage of the battery to further exploit the residual battery energy.
3. High switching frequency (> 1MHz) if the switching converter is adopted to reduce the
sizes of peripheral components such as the inductor and capacitor.

Table 4.3-2 DC/DC voltage conversion topology hierarchy ([58])

<table>
<thead>
<tr>
<th>Topology</th>
<th>Pros</th>
<th>Cons</th>
</tr>
</thead>
<tbody>
<tr>
<td>Linear regulator</td>
<td>Inexpensive, Very small, Low quiescent current, Low noise/EMI</td>
<td>Only steps down (Vout less than Vin)</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Inefficient at high input voltages</td>
</tr>
<tr>
<td>Buck converter</td>
<td>Lowest peak current, Only one switch voltage drop, Low-ripple current in output-filter capacitor, Simple inductor, Low switch-stress voltage</td>
<td>Only steps down (Vout less than Vin)</td>
</tr>
<tr>
<td></td>
<td></td>
<td>High-side switch</td>
</tr>
<tr>
<td>Boost converter</td>
<td>Low peak current, Low-side switch, Simple inductor, Low switch-stress voltage</td>
<td>Only steps up (Vout&gt;Vin)</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Output can't be completely turned off</td>
</tr>
<tr>
<td></td>
<td></td>
<td>No short-circuit protection</td>
</tr>
<tr>
<td>Charge pump</td>
<td>Inexpensive, Very small, Can boost or invert</td>
<td>Limited output power</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Limited range of input/output voltage ratio</td>
</tr>
</tbody>
</table>

The linear regulator can produce a stable output and is the simplest and most compact amongst all converter topologies. It will only work, however, when the input voltage is greater than the output, and the efficiency is poor from the voltage difference between the input and the output being larger. The voltage level of battery stacks drops as time goes by, and it will fail to power the linear regulator with a substantial amount of capacity still left and then wasted. Hence, the linear regulator does not seem to be a good topology for our application in terms of efficiency.

The charge pump converter is also simple and compact, only second to the linear regulator, in topology, but it is more versatile than the linear regulator in that it can be a boost converter, a
down converter, or a voltage inverter. However, its output power is limited and its output voltage will become a regulated one only if it is followed by a linear regulator, reducing the overall efficiency and adding complexity to the converter topology.

For the buck converter to be working the input voltage has to always be higher than the output one. This means that we have to put more batteries in series to generate a higher input voltage for the buck converter, which is unlikely in our case due to the unbalanced discharging in a series battery string.

Table 4.3-3 Main specifications for LM2623 boost converter of National Semiconductor ([64])

<table>
<thead>
<tr>
<th>Model Number</th>
<th>LM2623</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input Voltage Range, V</td>
<td>0.8 ~ 14</td>
</tr>
<tr>
<td>Minimum Startup Voltage, V</td>
<td>1.1</td>
</tr>
<tr>
<td>Output Voltage Range, V</td>
<td>1.24 ~ 14 (adjustable)</td>
</tr>
<tr>
<td>Internal Switch Current Rating, A</td>
<td>2.85</td>
</tr>
<tr>
<td>Maximum Output Current, A</td>
<td>2</td>
</tr>
<tr>
<td>Maximum Switching Frequency, MHz</td>
<td>2</td>
</tr>
<tr>
<td>Quiescent Current, µA</td>
<td>80</td>
</tr>
<tr>
<td>Shutdown Current, µA</td>
<td>&lt; 2</td>
</tr>
</tbody>
</table>

A boost converter and/or a buck/boost converter with high efficiency and low operating voltage seems to be more suitable for our application because they can fully exploit the useful voltage range of battery stacks and can function normally even when the voltage of battery stacks drops below the cutoff voltage, exploiting more energy remaining in the batteries.
Table 4.3-3 lists the main specifications for a switching boost converter of National Semiconductor ([64]) that can meet the input/output requirements of a power amplifier.

The minimum startup voltage for this converter is 1.1V and once it starts up the input voltage can drop to as low as 0.8V.

The above minimum input voltage for an LM2623 converter is only for a no-load condition. The minimum input voltage for a boost converter to maintain a specified output voltage while supplying a specified load current is dependent on the following equation:

\[ \eta(I_{in}V_{in}) = I_{out}V_{out} \]  \hspace{1cm} (4.4-1)

Wherein \( \eta \): Converter efficiency,

\( I_{in} \): input current,

\( V_{in} \): input voltage,

\( I_{out} \): Load current,

\( V_{out} \): Load voltage.

From (4.4-1) we can get:

\[ V_{in} = \frac{I_{out}V_{out}}{\eta I_{in}} \]  \hspace{1cm} (4.4-2)

In Fig. 4.4-1 a simplified diagram for a switching boost converter. The output voltage, Vout,
is brought down by a resistive divider and compared with a reference voltage to control the switching action of the switch. The current limit of the switch is an important specification for a boost converter. The input current, $I_{in}$, cannot exceed this current limit or the switch will be forced to turn off and the converter will cease operation. For specified output voltage and current, the higher the current limit of the switch, the lower the required input voltage as indicated by (4.4-2).

![Diagram of a switching boost converter](image)

**Fig. 4.4-1** Simplified diagram for a switching boost converter. The output voltage, $V_{out}$, is brought down by a resistive divider and compared with a reference voltage to control the switching action of the switch.

The switch limit current for LM2623 is 2.85A. For an output voltage 5V, an output current 1A and an average efficiency of 80%, it is likely to have a minimum input voltage from 2V to 3V.
CHAPTER 5
RESULT

5.1 Specifications of Cells Under Test

The battery that will meet our application should feature a low profile, a high volumetric energy density, a long shelf life, and the capability of delivering relatively high current.

Table 5.1-1 Specifications for the cells under test

<table>
<thead>
<tr>
<th>Technology</th>
<th>Zinc/Silver Oxide (Zn/Ag₂O)</th>
<th>Zinc/Silver Oxide (Zn/Ag₂O)</th>
<th>Lithium/Manganese Dioxide (Li/MnO₂)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Manufacturer</td>
<td>Maxell</td>
<td>Duracell</td>
<td>Panasonic</td>
</tr>
<tr>
<td>Model Number</td>
<td>SR1130SW</td>
<td>D389</td>
<td>CR2412</td>
</tr>
<tr>
<td>Size (Ø x H mm³)</td>
<td>11.6 x 3.05</td>
<td>11.6 x 3.05</td>
<td>24.5 x 1.2</td>
</tr>
<tr>
<td>Weight (g)</td>
<td>1.2</td>
<td>1.2</td>
<td>2</td>
</tr>
<tr>
<td>Rated Capacity (mAh)</td>
<td>79</td>
<td>70</td>
<td>100</td>
</tr>
<tr>
<td>Rated Discharge Current (μA)</td>
<td>100</td>
<td>110</td>
<td>200</td>
</tr>
<tr>
<td>Nominal Voltage (V)</td>
<td>1.55</td>
<td>1.55</td>
<td>3</td>
</tr>
<tr>
<td>Cutoff Voltage (V)</td>
<td>1.2</td>
<td>1.2</td>
<td>2</td>
</tr>
<tr>
<td>Actual Energy Density (Wh/l)</td>
<td>1193.4</td>
<td>1057.5</td>
<td>1666</td>
</tr>
<tr>
<td>Actual Energy Density (Wh/kg)</td>
<td>90</td>
<td>78.75</td>
<td>125</td>
</tr>
</tbody>
</table>

Most of the lithium-based batteries and the zinc/silver oxide battery of coin type do have these features, with the exception of a high current capability. The lithium/vanadium oxide battery, being only few of them, is capable of delivering high current at the price of high
procurement cost. Thus, the battery cells that we can choose to test at the current stage are restricted by their commercial availability.

The cells we chose to test were the silver oxide battery and lithium/manganese dioxide battery of the coin type. Table 5.1-1 lists the main specifications of the cells under test.

The rated discharge current in Table 5.1-1 is the best discharge condition recommended by the manufacturer to exploit the rated capacity, which is also the optimal capacity the battery could get. The batteries in the table are widely used in portable and small electronic applications, but they are not suitable for high rate applications. Fig. 5.1-1 and 5.1-2 are the measured load current and battery voltage versus capacity delivered in mAh for the two types of battery technologies, lithium/manganese dioxide and zinc/silver oxide.

Both cells are discharged via a 1Ω load resistor at room temperature and both reveal a total contrast in discharge behaviors. The lithium/manganese dioxide cell could deliver up to 1 amp of current at first, but the current level gradually drops with the voltage level. However, the current and voltage levels of the zinc/silver oxide cell remain quite stable till the cutoff voltage. If we consider the cutoff voltages claimed by both manufacturers as the end of battery life, the zinc/silver oxide cell delivers 37% of its rated capacity before its voltage drops down to 19% of its rated voltage while the lithium/manganese dioxide cell delivers only 4% of the rated capacity before the voltage drops down to 33% of its rated voltage. The zinc/silver oxide cell seems to deliver a large portion of its usable capacity before the cutoff voltage while the lithium/manganese dioxide cell does not.

Judging from the discharge behaviors, it is obvious that the zinc/silver oxide battery with a stable discharge profile is closer to the ideal battery than the lithium/manganese dioxide battery
with a dropping discharge profile. The power sources in our application will be used to power RF components such as low noise amplifier and power amplifier. These components need relatively accurate and stable voltage levels for proper operations, so the battery with a stable discharge profile, such as the zinc/silver oxide battery, is then preferable in our application. The following tests will be focused on the discharge behaviors of the zinc/silver oxide battery from a different manufacturer, Maxell SR1130SW, due to its higher energy density than that of Duracell D389.

Fig. 5.1-1 The measured load current and battery voltage versus capacity delivered in mAh for Panasonic CR2412 lithium/manganese dioxide battery. The cell is discharged via a 1Ω load resistor at room temperature. The straight line in the figure is the cutoff voltage claimed by the manufacturer.
Fig. 5.1-2 The measured load current and battery voltage versus capacity delivered in mAh for Duracell D389 zinc/silver oxide battery. The cell is discharged via a 1Ω load resistor at room temperature. The straight line in the figure is the cutoff voltage claimed by the manufacturer.
5.2 Constant Current Discharge Test for Maxell SR1130SW Zn/Ag₂O Battery

The battery cells are discharged under 7 current levels, 0.5, 1, 2, 5, 10, 20, and 40 mA, and at 2 temperatures, 21.5°C and 39.9°C. Fig. 5.2-1 is the measured voltage profile plotted against time in the logarithmic scale under seven constant discharge currents at room temperature (21.5°C).

![Silver Oxide Battery Discharged at 21.5°C](image)

Fig. 5.2-1 The measured voltage profiles plotted against time in the logarithmic scale under seven constant discharge currents at 21.5°C, 40mA, 20mA, 10mA, 5mA, 2mA, 1mA, and 0.5mA, from left to right.

As one can expect, the higher the discharge current the shorter the battery can last. The discharge profiles are stable for all the discharge currents except for those at higher discharge currents. For these, the battery voltage rises gradually and slightly from the beginning till the
middle of the discharge test, and then drops gradually to a plateau, remaining stable before
dropping under the cutoff voltage.

Fig. 5.2-2 is the measured voltage profile plotted against the capacity delivered by the
battery under the seven constant discharge currents at room temperature (21.5°C). It can be seen
from the figure that the higher the discharge current, the less the capacity the battery could deliver.
The battery becomes more inefficient in delivering energy as the discharge current increases.

Fig. 5.2-2 The measured voltage profiles plotted against the capacity delivered by the battery under seven
constant discharge currents at 21.5°C, 40mA, 20mA, 10mA, 5mA, 2mA, 1mA, and 0.5mA, from left to right.

Fig. 5.2-3 is the measured voltage profile plotted against time in the logarithmic scale under
seven constant discharge currents at an elevated temperature (39.9°C), while Fig. 5.2-4 is the
measured voltage profile plotted against the capacity delivered by the battery under the seven constant discharge currents at room temperature (39.9°C). They all show a similar trend when discharged at 21.5°C. The only difference is that the battery can deliver more capacity at an elevated temperature for each discharge current. Moreover, the capacity that the battery can deliver when discharged at 0.5mA is almost equal to the rated capacity.

![Silver Oxide Battery Discharged at 39.9°C](image)

**Fig. 5.2-3** The measured voltage profiles plotted against time in the logarithmic scale under seven constant discharge currents at 39.9°C, 40mA, 20mA, 10mA, 5mA, 2mA, 1mA, and 0.5mA, from left to right.

**Fig. 5.2-5** depicts the usable capacity that the battery can deliver when discharged under seven current levels, 40mA, 20mA, 10mA, 5mA, 2mA, 1mA, and 0.5mA, from right to left. The
The upper curve is discharged at 39.9°C while the lower one 21.5°C.

![Silver Oxide Battery Discharged at 40°C Voltage vs. Delivered Capacity](image)

Fig. 5.2-4 The voltage profiles plotted against the capacity delivered by the battery under seven constant discharge currents at 39.9°C, 40mA, 20mA, 10mA, 5mA, 2mA, 1mA, and 0.5mA, from left to right.

It seems that there’s a distinction between the low and high discharge currents. This distinction is particularly clear in the 21.5°C curve where the curve section smaller than 5 mA is not transitioned smoothly to the section greater than 5 mA. If we use the Peukert equation, $I^{n} = K$, to curve-fit the data, we get the corresponding coefficients, $n$ and $K$, as shown in Table 5.2-2. It should be noted that according to the discharge conditions, the data is divided into three groups for curving-fitting, (1) the data under all discharge conditions, (2) the data under small discharge...
conditions, 0.5 mA, 1 mA, and 2 mA, and (3) the data under large discharge conditions, 5 mA, 10 mA, 20 mA, and 40 mA.

Silver Oxide Battery
Delivered Capacity vs. Discharge Current
at 21.5°C and 39.9°C

Fig. 5.2-5 The usable capacity that the battery can deliver when discharged under seven current levels, 40 mA, 20 mA, 10 mA, 5 mA, 2 mA, 1 mA, and 0.5 mA, from right to left. The upper curve is discharged at 39.9°C while the lower one at 21.5°C.

From Table 5.2-1, it can be seen that there’s big difference on the power n between small discharge currents and the large discharge currents. It means that the capacity loss is greater at higher discharge current due to its greater power of n.

The Peukert constant K is dependent on temperature while the power n is not. The distinction between the low discharge currents and the large ones indicates that there are different
discharge mechanisms for both currents. Now the dependence of Peukert constant $K$ on temperature can be derived from the modified form incorporating the temperature dependence, $I^n t = K(1+\alpha T)$, as shown in Table 5.2-2.

Table 5.2-1 Curve-fitting coefficients of the Peukert equation, $I^n t = K$, for three groups of data, (1) the data under all discharge conditions, (2) the data under small discharge conditions, 0.5 mA, 1 mA, and 2 mA, and (3) the data under large discharge conditions, 5 mA, 10 mA, 20 mA, and 40 mA, at 21.5°C and 39.9°C.

<table>
<thead>
<tr>
<th>Curve-fitting for $I^n t = K$</th>
<th>21.5°C</th>
<th>39.9°C</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$n$</td>
<td>$K$</td>
</tr>
<tr>
<td>All Discharge Currents</td>
<td>1.372</td>
<td>68.885</td>
</tr>
<tr>
<td>Small Discharge Currents</td>
<td>1.186</td>
<td>69.347</td>
</tr>
<tr>
<td>Large Discharge Currents</td>
<td>1.43</td>
<td>80.499</td>
</tr>
</tbody>
</table>

Table 5.2-2 Curve-fitting parameters for modified Peukert equation incorporating temperature dependence, $I^n t = K(1+\alpha T)$, at small and large discharge currents

<table>
<thead>
<tr>
<th>Curve-fitting for $I^n t = K(1+\alpha T)$</th>
<th>21.5°C</th>
<th>39.9°C</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$n$</td>
<td>$K$</td>
</tr>
<tr>
<td>Small Discharge Currents (I $\leq$ 5 mA)</td>
<td>1.183</td>
<td>65.087</td>
</tr>
<tr>
<td>Large Discharge Currents (I $&gt;$ 5 mA)</td>
<td>1.438</td>
<td>42.111</td>
</tr>
</tbody>
</table>
5.3 Constant Current Discharge Test for Panasonic CR2412 Li/MnO$_2$ Battery

The battery cells are discharged under 7 current levels, 0.7, 1.1, 3.3, 5.6, 13.3, 27.8, and 58mA, and at 2 temperatures, 21.5°C and 39.9°C. Fig. 5.3-1 is the measured battery voltage profile plotted against time at four smaller discharge currents (0.7, 1.1, 3.3, and 5.6mA) at room temperature (21.5°C). The discharge times for higher current levels (13.3, 27.8, and 58mA) are too short to be shown in the figure.

![Lithium Manganese Dioxide Battery Discharged at 21.5°C](image)

Fig. 5.3-1 The measured voltage profiles against time of CR2412 battery discharged at 0.7mA, 1.1mA, 3.3mA, 5.6mA, from right to left, at 21.5°C.

The discharge profiles are not as flat as those of Zn/Ag$_2$O batteries. Fig. 5.3-2 is the measured
voltage profile plotted against the capacity delivered by the battery under the seven constant discharge currents at room temperature \((21.5^\circ\text{C})\). Fig. 5.3-3 is the measured battery voltage profile plotted against time at six smaller discharge currents \((0.7, 1.1, 3.3, 5.6, 13.3, \text{ and } 27.8\text{mA})\) at \(39.9^\circ\text{C}\). The discharge time for \(58\text{mA}\) at \(39.9^\circ\text{C}\) is too short to be shown in the figure. Fig. 5.3-4 is the measured voltage profile plotted against the capacity delivered by the battery under the seven constant discharge currents at \(39.9^\circ\text{C}\).

![Graph showing battery voltage profile](image)

**Lithium Manganese Dioxide Battery Discharged at 21.5°C**

**Voltage vs. Delivered Capacity**

Fig. 5.3-2 The measured voltage profiles plotted against time in the logarithmic scale under seven constant discharge currents at \(21.5^\circ\text{C}, 58\text{mA}, 27.8\text{mA}, 13.3\text{mA}, 5.6\text{mA}, 3.3\text{mA}, 1\text{mA}, \text{ and } 0.7\text{mA},\) from left to right.
Fig. 5.3-3 The measured voltage profiles against time of CR2412 battery discharged at 0.7mA, 1.1mA, 3.3mA, 5.6mA, 13.3mA, and 27.8mA from right to left, at 39.9°C.

Fig. 5.3-5 depicts the usable capacity that the battery can deliver when discharged under seven current levels, 58mA, 27.8mA, 13.3mA, 5.6mA, 3.3mA, 1mA, and 0.7mA, from right to left. The upper curve is discharged at 39.9°C while the lower one at 21.5°C. A distinction exists between the low and high discharge currents. This distinction is particularly clear in the 21.5°C curve where the curve section smaller than 13.3 mA is not transitioned smoothly to the section greater than 13.3 mA. If we use the Peukert equation, \( I^n \times t = K \), to curve-fit the data, we get the corresponding coefficients, \( n \) and \( K \), as shown in Table 5.3-1. It should be noted that according to the discharge conditions, the data is divided into three groups for curving-fitting, (1) the data
under all discharge conditions, (2) the data under small discharge conditions, 0.7mA, 1mA, 3.3mA, 5.6mA, and 13.3mA and (3) the data under large discharge conditions, 27.8mA and 58mA.

![Graph](image)

**Fig. 5.3-4** The measured voltage profiles plotted against time in the logarithmic scale under seven constant discharge currents at 39.9°C, 58mA, 27.8mA, 13.3mA, 5.6mA, 3.3mA, 1mA, and 0.7mA, from left to right.
Fig. 5.3-5 The usable capacity that the battery can deliver when discharged under seven current levels, 58mA, 27.8mA, 13.3mA, 5.6mA, 3.3mA, 1mA, and 0.7mA, from right to left. The upper curve is discharged at 39.9°C while the lower one 21.5°C.

The power n and the Peukert constant of CR2412 battery are all dependent on the temperature, unlike a SR1130SW battery in which case the power n is independent of the temperature.
Table 5.3-1 Curve-fitting coefficients of the Peukert equation, $I^{n}t = K$, for three groups of data, (1) the data under all discharge conditions, (2) the data under small discharge conditions, 0.7mA, 1mA, 3.3mA, 5.6mA, 13.3mA and (3) the data under large discharge conditions, 27.8mA and 58mA, at 21.5°C and 39.9°C.

<table>
<thead>
<tr>
<th>Curve-fitting for $I^n t = K$</th>
<th>21.5°C</th>
<th>39.9°C</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$n$</td>
<td>$K$</td>
</tr>
<tr>
<td>All Discharge Currents</td>
<td>2.168</td>
<td>101.629</td>
</tr>
<tr>
<td>Small Discharge Currents</td>
<td>1.757</td>
<td>75.240</td>
</tr>
<tr>
<td>Large Discharge Currents</td>
<td>3.508</td>
<td>9485.047</td>
</tr>
</tbody>
</table>
5.4 Comparisons

Fig. 5.4-1 lists the typical measured discharge curves for the Panasonic CR2412 Lithium/Manganese Dioxide battery and the Maxell SR1130SW Silver Oxide battery at two different temperatures.

Fig. 5.4-1 Typical measured discharge curves against time for the Panasonic CR2412 Lithium/Manganese Dioxide battery and the Maxell SR1130SW Silver Oxide battery at two different temperatures. The discharge curves stand for, from top to bottom, Panasonic CR2412 Lithium/Manganese Dioxide battery discharged at 0.7mA and 39.9°C, Panasonic CR2412 Lithium/Manganese Dioxide battery discharged at 0.7mA and 21.5°C, Maxell SR1130SW Silver Oxide battery discharged at 0.5mA and 39.9°C, and Maxell SR1130SW Silver Oxide battery discharged at 0.5mA and 21.5°C.
It is obvious that the discharge profiles for the Maxell SR1130SW Silver Oxide battery remain flat till the cutoff voltage. Though the cutoff voltage for the Maxell SR1130SW Silver Oxide battery is 1.2V as claimed by the manufacturer, the actual battery voltage is kept relatively constant and above 1.5V throughout the entire battery life, a feature resembling the ideal battery.

While for the Panasonic CR2412 Lithium/Manganese Dioxide battery, the discharge curves are dropping down to the cutoff voltage, 2V. The cutoff voltage for the Panasonic CR2412 Lithium/Manganese Dioxide battery is 2V as claimed by the manufacturer, but from the discharge profile it seems that after the battery voltage drops below 2.5V, the discharge curve plunges downward even more sharply.

If we define the internal resistance of the battery as following:

\[
R = \frac{V_{oc} - V_{cc}}{I}
\]  

(5.1)

Where Voc: open-circuit voltage of a battery before discharge

Vcc: closed-circuit voltage during discharge

I: discharge current

The resistance of the Maxell SR1130SW Silver Oxide battery during the constant current discharge varies little throughout the discharge time span and increases suddenly at the end of discharge. The resistance of the Maxell SR1130SW Silver Oxide battery during the constant current discharge increases gradually towards the end of life.
Fig. 5.4-2 depicts Normalized power delivered plotted against normalized energy delivered for the Panasonic CR2412 Lithium/Manganese Dioxide battery and the Maxell SR1130SW Silver Oxide battery at two different temperatures.

The normalized power is defined by the ratio of actual power to the rated power of the battery, and the normalized energy is defined by the ratio of actual energy delivered to the rated energy of the battery. The rated energy of a battery is the product of the capacity and the rated voltage, and the rated power is the product of the 1C rate and the rated voltage. For example, the Panasonic CR2412 Lithium/Manganese Dioxide battery is rated at 3V and 100mAh, so its rated energy is 3V*100mAh = 0.3Watts-hours and its rated power is 100mA*3V = 0.3Watts. For an ideal battery, it should always deliver the rated energy under any discharge power rates. However, as mentioned before, the actual battery becomes more inefficient from an energy point of view as the discharge current and thus power increases.

In Fig. 5.4-2, for the same normalized energy delivered by the battery, the Maxell SR1130SW Silver Oxide battery can deliver higher power than the Panasonic CR2412 Lithium/Manganese Dioxide battery. For delivering the same normalized power, the Maxell SR1130SW Silver Oxide battery can deliver more energy than does the Panasonic CR2412 Lithium/Manganese Dioxide battery.

Although the theoretical energy density of the Lithium/Manganese Dioxide battery is higher than that of the Silver Oxide battery, the actual energy available for the Lithium/Manganese Dioxide battery turns out to be less than that of the Silver Oxide battery as the discharge current/power increases. Hence, the Silver Oxide battery is more suitable for high power application than the Lithium/Manganese Dioxide battery.
Fig. 5.4-2 Normalized power delivered plotted against normalized energy delivered for the Panasonic CR2412 Lithium/Manganese Dioxide battery and the Maxell SR1130SW Silver Oxide battery at two different temperatures. The curves stand for, from bottom to top, the Panasonic CR2412 Lithium/Manganese Dioxide battery discharged at 21.5°C, the Panasonic CR2412 Lithium/Manganese Dioxide battery discharged at 39.9°C, the Maxell SR1130SW Silver Oxide battery discharged 21.5°C, and the Maxell SR1130SW Silver Oxide battery discharged at 39.9°C. The normalized power is defined by the ratio of actual power to the rated power of the battery, and the normalized energy is defined by the ratio of actual energy delivered to the rated energy of the battery.
From the test results for small discharge current, it is shown that a silver oxide cell would deliver 99% of its total capacity before the cell voltage drops under 1.5V. Since the discharge current of battery cells in our application will be limited to a small value within the optimal discharge current range, we can assume that the cut-off voltage for a silver oxide cell discharged under low current level is 1.5V instead of 1.2V as claimed by the manufacturers. As for the lithium manganese dioxide cell, its discharge profile is not as flat as that of the silver oxide cell even under a small discharge current. The useful voltage range for a lithium manganese dioxide cell is still from 2V to 3V as claimed by the manufacturers.
5.5 Performance Evaluation for the Boost Converter with the Supercapacitor

The converter under test is LM2623 DC-DC boost converter. Its specification is listed in Table 5.5-1.

Table 5.5-1 Main specifications for LM2623 boost converter of National Semiconductor ([64])

<table>
<thead>
<tr>
<th>Specification</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Model Number</td>
<td>LM2623</td>
</tr>
<tr>
<td>Input Voltage Range, V</td>
<td>0.8 ~ 14</td>
</tr>
<tr>
<td>Minimum Startup Voltage, V</td>
<td>1.1</td>
</tr>
<tr>
<td>Output Voltage Range, V</td>
<td>1.24 ~ 14 (adjustable)</td>
</tr>
<tr>
<td>Internal Switch Current Rating, A</td>
<td>2.85</td>
</tr>
<tr>
<td>Maximum Output Current, A</td>
<td>2</td>
</tr>
<tr>
<td>Maximum Switching Frequency, MHz</td>
<td>2</td>
</tr>
<tr>
<td>Quiescent Current, μA</td>
<td>80</td>
</tr>
<tr>
<td>Shutdown Current, μA</td>
<td>&lt; 2</td>
</tr>
</tbody>
</table>

The schematic of the performance test circuitry for the DC-DC boost converter with the supercapacitor is shown in Fig. 5.5-1. The capacitance of supercapacitor is 0.12F.

The output of the converter is connected to a programmable electronic load operating under the dynamic constant-current mode. The duration and frequency of the load current pulses can be programmed via the keypad at the front panel of the electronic load. The power source charging up the supercapacitor could be either battery cells or a DC power supply. In both cases, a current-limiting resistor is added between the power source and the supercapacitor to mimic the nature of the limited current drain of batteries. Two current sensing resistors of small resistance,
0.01Ω or 0.05Ω, are inserted to sense the input and output currents. When the input voltage is below the rated voltage, 2.25 Volts, of single cell of the supercapacitor in the case that one battery cell is used, only one cell of the supercapacitor is connected and no balancing resistors are needed. The load current, $I_{out}$, is the controlled parameter with the presettable current level and pulse duration. The voltage of the supercapacitor, $V_{in}$, the input current, $I_{in}$, the output voltage, $V_{out}$ and $I_{out}$ are then recorded. The sampling time is $10 \mu S$.

![Fig. 5.5-1 The schematic of the performance test circuitry for the DC-DC boost converter with the supercapacitor. The output of the converter is connected to a programmable electronic load operating under the dynamic constant-current mode with the presettable current level and pulse duration. The power source charging up the supercapacitor could be either battery cells or a DC power supply. In both cases, a current-limiting resistor is added between the power source and the supercapacitor to mimic the nature of the limited current drain of batteries. Two current sensing resistors of small resistance, 0.01Ω or 0.05Ω, are inserted to sense the input and output currents, $I_{in}$ and $I_{out}$. Vsource in Fig. 5.5-1 is voltage of input power source. Fig. 5.5-2 is the measured profiles.](image)
for Vin, Iin, Vout, and Iout when Vsource is set to 1.5V, equivalent to the voltage of one cell of silver oxide battery, and Iout 1A with 40 μS of duration. When the load is applied, the output voltage of the converter remains within the operating ranges of power amplifier, which are 4.75V minimum and 5.25V maximum.

![Diagram](image)

Fig. 5.5-2 The measured profiles for Vin, Iin, Vout, and Iout when Vsource is set to 1.5V. When the load is applied, the output voltage of the converter remains within the operating ranges of the power amplifier, which are 4.75V minimum and 5.25V maximum.

1A and 40μS pulse is the limit that LM2623 boost converter can generate out of an input source voltage of 1.5V. If we increase the load current or elongate the pulse duration, the output voltage of the converter will fall under the minimum operating voltage of power amplifier and
fail to drive it when the load is applied.

Fig. 5.5-3 are the measured profiles for Vin, lin, Vout, and Iout when Vsource is set to 3V, equivalent to the voltage of two cells of silver oxide batteries in series, and Iout 1.5A with 30 μS of duration.

Fig. 5.5-4 are the measured profiles for Vin, lin, Vout, and Iout when Vsource is set to 3V, equivalent to the voltage of two cells of silver oxide batteries in series, and Iout 1.0A with 100 μS of duration.

It is shown that for the input voltage of 3V, the boost converter will be able to supply a 1.5A pulse for a very short duration, 30 μS. From the input current profile, it also indicates that the converter can barely respond to the load in such a short period of time, so the input current hasn’t risen enough to match the load before it is cut off. When the load requirement is reduced to 1A, the output voltage of the converter is able to sustain longer pulse duration, 100 μS, without falling out of the operating voltage range of power amplifier.

Fig. 5.5-5 are the measured profiles for Vin, lin, Vout, and Iout when Vsource is set to 4.5V, equivalent to the voltage of three cells of silver oxide batteries in series, and Iout 1.95A with 150 μS of duration.

Fig. 5.5-6 are the measured profiles for Vin, lin, Vout, and Iout when Vsource is set to 4.5V, equivalent to the voltage of three cells of silver oxide batteries in series, and Iout 1.5A with 200 μS of duration.

As indicated by the Fig. 5.5-5 and Fig. 5.5-6, the converter can generate higher current level and longer pulse duration when the input source is 3 silver oxide batteries with a higher source voltage, 4.5V.
Fig. 5.5-3 The measured profiles for $V_{in}$, $I_{in}$, $V_{out}$, and $I_{out}$ when $V_{source}$ is set to 3V. When the load is applied, the output voltage of the converter remains within the operating ranges of the power amplifier, which are 4.75V minimum and 5.25V maximum.
Fig. 5.5-4 The measured profiles for Vin, lin, Vout, and Iout when Vsource is set to 3V. When the load is applied, the output voltage of the converter remains within the operating ranges of the power amplifier, which are 4.75V minimum and 5.25V maximum.
Fig. 5.5-5 The measured profiles for Vin, Iin, Vout, and Iout when Vsource is set to 4.5V. When the load is applied, the output voltage of the converter remains within the operating ranges of the power amplifier, which are 4.75V minimum and 5.25V maximum.
Fig. 5.5-6 The measured profiles for Vin, Iin, Vout, and Iout when Vsource is set to 4.5V. When the load is applied, the output voltage of the converter remains within the operating ranges of the power amplifier, which are 4.75V minimum and 5.25V maximum.
5.6 Leakage Test for Supercapacitor

The supercapacitor under test is GW209D of Cap-XX. The specification of GW209D is listed in Table 5.6-1.

Table 5.6-1 Specification for Gw209D supercapacitor of Cap-XX

<table>
<thead>
<tr>
<th>Capacitor Technology</th>
<th>Super Capacitor</th>
</tr>
</thead>
<tbody>
<tr>
<td>Manufacturer</td>
<td>Cap-XX</td>
</tr>
<tr>
<td>Manufacturer's Model Number</td>
<td>GW 209D</td>
</tr>
<tr>
<td>Capacitance (F)</td>
<td>0.12</td>
</tr>
<tr>
<td>Nominal Voltage Rating (V)</td>
<td>4.5</td>
</tr>
<tr>
<td>Maximum Voltage Rating (V)</td>
<td>5</td>
</tr>
<tr>
<td>ESR (Ω) at 100 kHz</td>
<td>0.09</td>
</tr>
<tr>
<td>Leakage Current (µA)</td>
<td>&lt; 2</td>
</tr>
<tr>
<td>Dimensions (L x W x H mm³)</td>
<td>28.5 x 17 x 2.06</td>
</tr>
<tr>
<td>Weight (gm)</td>
<td>0.6</td>
</tr>
<tr>
<td>Maximum Pulse Current (A)</td>
<td>30</td>
</tr>
</tbody>
</table>

The GW209D supercapacitor is actually composed of two cells in series with a rated voltage 2.25V each to increase the voltage rating to 4.5V. When charging or discharging the supercapacitor, two balancing resistors of equal resistance should be in parallel with each cell to balance the voltage of each cell and prevent them from being over-charged. The resistance of the balancing resistors, as suggested by the manufacturer, can be any value from 39 KΩ to 100 KΩ. Before testing, the GW209D supercapacitor is charged up at 4.5 Volts by a DC power supply and balanced by two balancing resistors for three days. The power supply and the balancing resistors are then removed and the voltage decay of the supercapacitor under test is recorded.
The measured voltage decay and the equivalent leakage current versus time for the GW209D supercapacitor at room temperature is shown in Fig. 5.6-1.

The equivalent leakage current is defined as \( \frac{C \times \Delta V}{\Delta t} \), wherein \( C \) is the nominal capacitance of the supercapacitor, \( \Delta V \) the difference between adjacent voltage measurements, and \( \Delta t \) the time difference between adjacent measurements.

![Graph showing measured voltage decay and equivalent leakage current versus time for the GW209D supercapacitor.](image)

Fig. 5.6-1 The measured voltage decay and the equivalent leakage current versus time for the GW209D supercapacitor at room temperature.

The equivalent leakage current drops down to 1 \( \mu \)A after 3 days and 0.3 \( \mu \)A after 10 days. And the equivalent leakage current continues to go down as time goes on.

The measured voltage decay and the equivalent leakage current versus time for the
GW209D supercapacitor at 40°C is shown in Fig. 5.6-2. The comparison of equivalent leakage current between room temperature, 25°C, and 40°C is shown in Fig. 5.6-3. The equivalent leakage current at 40°C is about 1.5 times higher than that at room temperature, but the difference is diminishing as the test time drags on as shown in Fig. 5.6-3.

![Graph](image)

Fig. 5.6-2 The measured voltage decay and the equivalent leakage current versus time for the GW209D supercapacitor at 40°C.
There is a second way to measure the leakage current using only one supercapacitor cell. By doing this, we can eliminate the involvement of balancing circuitry and can focus on the leakage current purely from the self-discharge of the supercapacitor. The cell has been charged up at 1.5V for 3 days via a 10KΩ charging resistor from a DC power supply. There is a great possibility that we will use two silver oxide cells in series to charge the whole supercapacitor pack at 3V. One cell at 1.5V means two cells at 3V in real application. When two supercapacitor cells are in series, their leakage current should be equal to that of one cell.

The voltage drop across the charging resistor is then recorded and converted into leakage current. The leakage current is monitored for a period of time to see whether it is settled down or not. If the leakage current is settled, the voltage setting of the power supply charging the
supercapacitor cell is brought down from 1.5V to 1V and then up to 1.5V again. This process is to monitor how the leakage current of the supercapacitor will change after a severe discharge. This forced discharge process is much more severe than that in our real application, because the worst voltage drop in our application will not be greater than 0.01V for the 0.12F supercapacitor and discharge duration no longer than 55mS.

Fig. 5.6-4 is the measured leakage current of a single supercapacitor cell over time at room temperature. The cell has been charged up at 1.5V for 3 days via a 10KΩ charging resistor from a DC power supply. Before time zero, the leakage current is monitored for a period of time. At time zero, the voltage setting of the power supply charging the supercapacitor cell is brought down from 1.5V to 1V and then up to 1.5V again. It can be shown in the figure that the leakage current is settled around 0.22 μA before the discharge and the leakage current doesn’t change after discharge and still settles to around 0.22 μA like the one before the severe discharge.

Fig. 5.6-5 is the measured leakage current of a single supercapacitor cell over time at 40°C. The cell has been charged up at 1.5V for 3 days via a 10KΩ charging resistor from a DC power supply. Before time zero, the leakage current is monitored for a period of time. At time zero, the voltage setting of the power supply charging the supercapacitor cell is brought down from 1.5V to 1V and then up to 1.5V again. It can be shown in the figure that the leakage current is settled to around 0.45 μA before the discharge and the leakage current doesn’t change after discharge and still settles to around 0.45 μA like the one before the severe discharge.
Fig. 5.6-4 The measured leakage current of a single supercapacitor cell over time at room temperature. The cell has been charged up at 1.5V for 3 days via a 10KΩ charging resistor from a DC power supply. Before time zero, the leakage current is monitored for a period of time. At time zero, the voltage setting of the power supply charging the supercapacitor cell is brought down from 1.5V to 1V and then up to 1.5V again. It can be shown in the figure that the leakage current is settled around 0.22 μA before the discharge and the leakage current doesn’t change after discharge and still settles to around 0.22 μA like the one before the severe discharge.
Fig. 5.6-5 The measured leakage current of a single supercapacitor cell over time at 40°C. The cell has been charged up at 1.5V for 3 days via a 10KΩ charging resistor from a DC power supply. Before time zero, the leakage current is monitored for a period of time. At time zero, the voltage setting of the power supply charging the supercapacitor cell is brought down from 1.5V to 1V and then up to 1.5V again. It can be shown in the figure that the leakage current is settled to around 0.45 µA before the discharge and the leakage current doesn't change after discharge and still settles to around 0.45 µA like the one before the severe discharge.
CHAPTER 6
CONCLUSION

6.1 Proposed Battery/Capacitor Combination

From the battery test results we know that there’s a current range within which the battery can be discharged to deliver the optimal capacity. Beyond that current range, the battery will incur capacity loss. If we want to exploit as much energy as possible from a battery, we have to make sure that it is always discharged within this optimal current range.

The optimal current range for the zinc/silver oxide battery of the test is less than 500 $\mu$A from the experiment result or 100 $\mu$A as recommended by the manufacturer. However, these current ratings are way below our application requirements, which is above 1 Amp level for the power amplifier. We should devise a strategy to protect our batteries from a high rate discharge and still be able to meet the load requirements.

The proposed hybrid power system is modified from Purdy [37] and redrawn in Fig. 6.1-1 with the addition of a decoupling resistor, $r$. This topology is used to enhance the output current capability and thus the output power of the battery by adding a parallel capacitor.

The role of the resistor $r$ is crucial in that its resistance has to be large enough to limit the current drain of the battery to the optimal current under any circumstances, so that the battery can deliver the optimal capacity. The capacitor can supply high current when the load demands while the battery is restricted by the decoupling resistor $r$ only to supply the optimal current and recharges the capacitor when the load idles.
Fig. 6.1-1 Hybrid capacitor/battery model proposed Purdy [37], wherein (1) $V_0$: load voltage, (2) $V_B$: battery voltage, (3) $I_0$: load current, (4) $I_B$: battery current, (5) $I_C$: capacitor current, (6) $R_B$: battery resistance, (7) $R_C$: equivalent capacitor resistance, (8) CSC: capacitance, (9) $r$: decoupling resistor.

The equations governing the equivalent circuit in Fig. 6.1-1 can be derived as following:

$$I_0 = I_C + I_B$$  \hspace{1cm} (6-1)

$$V_0 = V_B - I_B(R_B + r) = \left[ V_B - \frac{1}{C_{SC}} \int I_C dt \right] - I_C R_C$$  \hspace{1cm} (6-2)

$$V_B - V_0 = \Delta V = \frac{I_0(R_B + r)R_C}{(R_B + r) + R_C} + \frac{I_0(R_B + r)T}{C_{SC}}$$  \hspace{1cm} (6-3)

Wherein

$V_0$: load voltage
\( V_B \): battery voltage

\( I_0 \): load current

\( I_B \): battery current

\( I_C \): capacitor current

\( R_B \): battery resistance

\( R_C \): equivalent capacitor resistance

\( C_{SC} \): capacitance

\( r \): decoupling resistance

Equation (6-3) is derived under the assumption that the current of the capacitor during discharge remains constant. The first term in (6-3) is the voltage drop due to the ohmic resistances, and the second term is the voltage drop due to the discharge of the capacitor. The ratio of the current delivered by the battery (\( I_B \)) to that by the capacitor (\( I_C \)) can be derived as follows:

\[
\frac{I_C}{I_B} = \frac{I_0 - I_B}{I_B} = \frac{R_B + r}{R_C + \frac{T}{C_{SC}}}
\]

We want the ratio in (6-4) as great as possible so that the capacitor can supply the largest portion of the load current without over-burdening the battery. It can be seen that if the pulse duration is kept short and/or the capacitance is large, the term \( T/C_{SC} \) is negligible and the ratio of the discharge current would be determined by the internal resistances between the battery and the
capacitor. It is also necessary to have a capacitor with low internal impedance, to further enhance the current-sharing capability of the capacitor.

There are three kinds of load requirements in our application:

1. High voltage (~5V), high current (> 1A), short pulse (~100 μS), and low duty cycle (once a day) load for power amplifier,

2. Medium voltage (~3V), medium current (10~20mA), long pulse (tens of milliseconds) and high duty cycle load for low noise amplifier,

3. Medium voltage (< 3V), low current (< 2 μA most of the time), constant-running load for the microcontroller.

We certainly need the assistance of the capacitor for the first type of load. For the second type of load, even the battery itself can supply such current the usable capacity of the battery will still be diminished. So we still need the capacitor to assist supplying the second type of load.

As for the third type of load, since the microcontroller will draw more current, 1mA to 2mA, when power amplifier or low noise amplifier become active, a small capacitor is still needed to assist the battery supplying such loads.

The required capacitance Csc and the current-limiting resistance r in (6-4) depend on the voltage of power source, VB in Fig. 6.1-1, and the voltage drop of the capacitor during discharge. The voltage of power source is based on the battery technology and how many cells are connected in series. The voltage drop of the capacitor during discharge is dependent on the converter topology, namely, the minimum operating voltage of the converter for maintaining a
specified output current and voltage.

6.2 Proposed Battery Technology and Cell Combination

In this section we will decide the battery technology and how many battery cells connected in series based on that battery technology. Such a decision will most certainly affect the corresponding converter topology.

One cell of silver oxide battery will be counted out because the voltage, 1.55V, is too low to drive the LM2623 boost converter to produce the required input voltage and current for a power amplifier. Furthermore, the voltage is also so low for driving a low noise amplifier and microcontroller that another two boost converters will be needed.

The battery technologies in comparison are silver oxide and lithium manganese dioxide. The discharge profile for the silver oxide cell is almost flat if discharged at a small current. From the test results for the small discharge current, it is shown that a silver oxide cell would deliver 99% of its total capacity before the cell voltage drops under 1.5V. Since the discharge current of battery cells in our application will be limited to a small value within the optimal discharge current range, we can assume that the cut-off voltage for a silver oxide cell discharged under low current level is 1.5V instead of 1.2V as claimed by the manufacturers. As for the lithium manganese dioxide cell, its discharge profile is not as flat as that of the silver oxide cell even under a small discharge current. The useful voltage range for a lithium manganese dioxide cell is still from 2V to 3V as claimed by the manufacturers.

Fig. 6.2-1 is the possible cell combinations and the corresponding useful voltage ranges. The useful voltage range is defined as from the nominal voltage to the cut-off voltage as claimed
by manufacturers. However, the cut-off voltage for the silver oxide cell can be viewed as 1.5V based on the aforementioned reason.

The output voltage for LM2623 boost converter is 5V. The minimum input voltage for the LM2623 boost converter to supply 1A and 100uS pulse at 5V is 3V according to the test results. The input voltage range for a low noise amplifier is from 2.75V to 3.25V. When the battery voltage is lower than the required input voltage of a power amplifier, we need a boost converter; when the battery voltage is higher than that we need a buck one. A buck/boost converter is necessary should the required output voltage lie somewhere within the useful range of battery voltage. The description in the brackets beside each cell combination is the necessary converter topology to drive a power amplifier and a low noise amplifier.

1 cell of lithium manganese dioxide battery is not enough to drive a power amplifier as the cell voltage goes down below 3V. Cell combination 2, 2 cells of silver oxide battery, involves only one boost converter in the converter topology due to the unique discharge profile of a single silver oxide battery delivering most of its capacity above 1.5V under small discharge current.
Fig. 6.2-1 Possible cell combinations and the corresponding useful voltage ranges. The useful voltage range is defined as from the nominal voltage to the cut-off voltage as claimed by the manufacturers. The output voltage for the LM2623 boost converter is 5V. The minimum input voltage for the LM2623 boost converter to supply 1A and 100uS pulse at 5V is 3V according to the test results. The input voltage range for a low noise amplifier is from 2.75V to 3.25V.

A proposed converter topology based on the cell combination 2 in Fig. 6.2-1 is shown in Fig. 6.2-2. Two silver oxide battery cells (2 SR cells) in series can generate a useful voltage range from 3V to 3.1V to directly drive the microcontroller (uC). The battery stack is also connected in
parallel with a big capacitor via a current limiting resistor to restrain the discharge current from the battery stack to the capacitor. The capacitor is used as an input power source for the microcontroller, the low noise amplifier and for the boost converter to bring the input voltage up to 5V to drive the power amplifier when activated. The always-running microcontroller controls the activations and shutdowns of the low noise amplifier, boost converter, and power amplifier.

![Proposed converter topology based on the cell combination 2 in Fig. 6.2-1. Two silver oxide battery cells (2 SR cells) in series can generate a useful voltage range from 3V to 3.1V to directly drive the microcontroller (μC) and low noise amplifier (LNA). The battery stack is also connected in parallel with a big capacitor via a current limiting resistor to restrain the discharge current from battery stack to capacitor. All electronic components have built-in shutdown functionality. When entering the shutdown mode as commanded by the microcontroller, they only consume a current of microampere level.](image-url)

All electronic components have built-in shutdown functionality. When entering the shutdown mode as commanded by the microcontroller, they only consume a current of microampere level.
6.3 Proposed Converter Topology

In this section we will determine the details in the proposed converter topology in Fig. 6.2-2. There are two important parameters in the proposed converter topology to be determined. One is the capacitance of the capacitor, and the other is the resistance of the current-limiting resistor.

The capacitor is used to supply a relatively long current pulse, 18mA and 55mS, demanded by the low noise amplifier upon activation. The initial voltage of the capacitor is 3V and cannot drop down under 2.75V, the minimum input voltage of the low noise amplifier, during the entire period of the activation of the amplifier. The minimum required capacitance is obtained as follows:

\[ C_{\text{min}} = \frac{I \times \Delta t}{V_{\text{init}} - V_f} \]  \hspace{1cm} (6-5)

Wherein,

I: discharge current of capacitor

\( \Delta t \): discharge period of capacitor

V\text{init}: initial voltage of capacitor before discharge

V\text{f}: final voltage of capacitor right after discharge

In (6-5), we set I to be 18mA, \( \Delta t \) to be 55mS, V\text{init} to 3V, and V\text{f} to be 2.75V and then we get the minimum required capacitance of 3960 \( \mu \)F. Such capacitance is too large in terms of physical size for a conventional capacitor technology such as ceramic capacitors, and is too leaky for tantalum capacitors. With a capacitance of 3960 \( \mu \)F, the leakage current can be as high as
0.01C*V=100 \mu A for tantalum capacitors. The remaining charge in this tantalum capacitor of 3960 \mu F at 2.75V after powering a low noise amplifier would be completely drained out and wasted within 2 minutes. Thus, the capacitor technology in need for our application is one with a high volumetric efficiency and an extremely low leakage current.

The supercapacitor GW209D we’ve tested has a capacitance of 0.12F, an ESR of 0.09\Omega and the leakage current is below 2\mu A even at 40^\circ C. This supercapacitor is the smallest in capacitance and physical size and lowest in leakage current commercially available in the current market. The capacitance of the supercapacitor seems to be too high for our application. However, the greater the capacitance the lower the voltage drop of the capacitor will be during discharge, leading to an extended battery life. The worst voltage drop of this supercapacitor will be down to a negligible 0.01V right after the activation of a low noise amplifier. With this voltage drop between the battery stack and the supercapacitor, the resistance of the current-limiting resistor is 20\Omega if the discharge current of the battery stack is limited to 500 \mu A for fully exploiting the capacity of the battery. For safety reason, we may double the resistance to 40\Omega. The time constant of the supercapacitor will now be less than 10 seconds, a time period much shorter than the period of the low noise amplifier and certainly power amplifier leading to a virtually full-charging of the supercapacitor before its next discharge.

The problem with the supercapacitor is that the low voltage rating (2.25V) makes inevitable the stacking of multiple supercapacitor cells in series and hence the involvement of balancing circuitry. The balancing circuitry can be passive, using resistors of equal resistance in parallel with each supercapacitor cell, or active, using a voltage comparator to detect voltage difference between cells and then close a switch to inject current into one supercapacitor cell. The active
balancing is preferred in that it only consumes current when the voltage difference between cells occurs while passive balancing always consumes current, but the passive balancing is much simpler than the active counterpart in implementation by using only resistors.

Fig. 6.3-1 Detailed converter topology with two silver oxide battery cells in series. Each battery cell with a current-limiting resistor of 40 Ω can serve to balance each supercapacitor cell.

However, with two silver oxide battery cells in series, each battery cell with a current-limiting resistor can serve to balance each supercapacitor cell as shown in Fig. 6.3-1. The proposed two-battery cells combination can obviate the need for balancing circuitry and the power for maintaining its operation, whether active or passive,
6.4 Service Life

The goal of the power management is for the tag to last for 3 years. The energy budget bounded by the available board volume is 480mAh. However, the battery capacity will diminish as time passes due to self-discharge.

![Retained Capacity over years at 40°C](image)

Fig. 6.4-1 The retained capacity in the percentage of fresh capacity over time for a silver oxide battery and a lithium manganese dioxide battery at 40°C (Courtesy of Energizer Holdings, Inc. [66])

The retained capacity in the percentage of fresh capacity over time for a silver oxide battery and a lithium manganese dioxide battery at 40°C is shown in Fig. 6.4-1 [66]. The lithium
manganese dioxide battery has a lower self-discharge than the silver oxide battery. Such a difference becomes more prominent as the time period extends longer. For 3 years time as in our project goal, the difference is not that great as shown in Fig. 6.4-1. We will adhere to a silver oxide battery as our proposed battery technology.

Table 6.4-1. Detailed power requirements and activation durations for components in the proposed topology

<table>
<thead>
<tr>
<th>Components</th>
<th>Voltage, V</th>
<th>Current</th>
<th>Active Duration/ Duration between adjacent activations</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power Amplifier (PA) - Active</td>
<td>5</td>
<td>1A</td>
<td>100 μS/24hrs</td>
</tr>
<tr>
<td>Power Amplifier (PA) - Shutdown</td>
<td>5</td>
<td>1uA</td>
<td>—</td>
</tr>
<tr>
<td>Converter - Active</td>
<td>3</td>
<td>80uA</td>
<td>1mS(startup) + 100uS/24hrs</td>
</tr>
<tr>
<td>Converter - Shutdown</td>
<td>2.5</td>
<td>2uA</td>
<td>—</td>
</tr>
<tr>
<td>Low Noise Amplifier (LNA) - Active</td>
<td>3</td>
<td>18</td>
<td>55mS/to be calculated</td>
</tr>
<tr>
<td>Low Noise Amplifier (LNA) - Shutdown</td>
<td>3</td>
<td>0.3uA</td>
<td>—</td>
</tr>
<tr>
<td>Microcontroller - PA + Converter becomes active</td>
<td>3</td>
<td>2.5mA</td>
<td>1mS(startup) + 100 μS/24hrs</td>
</tr>
<tr>
<td>Microcontroller - LNA becomes active</td>
<td>3</td>
<td>1mA</td>
<td>1mS(startup) + 55mS/to be calculated</td>
</tr>
<tr>
<td>Microcontroller - Shutdown</td>
<td>3</td>
<td>1.8uA</td>
<td>—</td>
</tr>
</tbody>
</table>

Ideally, the low noise amplifier will remain active all the time in order not to miss any interrogation signals sent out from the base station. However, the limited energy budget cannot afford its constant-running operation. To compromise, the low noise amplifier has to be turned on for a certain period of time for listening and then turned to save power. The frequency of activation for the low noise amplifier has to be as high as possible, or namely, the duration between adjacent activations for the low noise amplifier has to be as short as possible, to increase the possibility of catching the interrogation signals while not too high to overburden the energy.
Fig. 6.4-2 Energy flow diagram for our proposed battery/supercapacitor system

Fig. 6.4-2 depicts the energy flow diagram for our proposed battery/supercapacitor system. The battery/supercapacitor system can be treated as a hybrid power source. In this system, the total energy available is \( Q \times V \). \( Q \) is the nominal capacity of the battery, and \( V \) is the battery voltage. It is sure that the available energy is equal to the nominal one because the battery is always discharged under the optimal current level by the restriction of the limiting resistor \( r \).

The total energy of this hybrid power source should give away to the following constituents:

1. Maintaining component’s operation when they are in active, startup, and shutdown modes as listed in Table 6.4-1 and illustrated in Fig. 6.4-3
2. Leakage of the supercapacitor and self-discharge of battery cells.

3. Discharging loss of the supercapacitor via its ESR and charging loss via the impedance of the battery, the current-limiting resistor, and ESR of supercapacitor.

![Diagram showing current levels in shutdown, startup, and active modes for components in the proposed converter topology.](image)

Fig. 6.4-3 Current levels in shutdown, startup, and active modes for components in the proposed converter topology.

In Fig. 6.4-3, the microcontroller has two different active current levels, one is 1mA upon the activation of low noise amplifier, and the other is 2.5mA upon the activations of both power amplifier and voltage converter. However, every electronic component has a startup or warm-up
period during the transition from shutdown mode to active mode. The longest transition time in the proposed converter topology is 1mS. The microcontroller has to become active during the startup of other electronic components, so its activation period is one startup period, 1mS, longer than the activation period of other electronic components. The voltage converter has to work in synchronization with the power amplifier to drive it upon activation. And the voltage converter also has a startup period, 1mS, during the transition from shutdown mode to active mode.

Fig. 6.4-4 The calculated energy consumption breakdown for the proposed converter topology of Fig. 6.3-1 operating for 3 years at 40°C.

The leakage current of the supercapacitor in this converter topology is well below 0.5 μA at 40°C. However, we will use 0.5 μA as the worst case scenario in the leakage loss of the supercapacitor.
The minimum duration between adjacent activations for the low noise amplifier can be calculated based on the energy flow diagram. For 3 year of operation at 40°C, the minimum duration between adjacent activations for the low noise amplifier is 1.8 minutes.

The calculated energy consumption breakdown for the proposed converter topology of Fig. 6.3-1 operating for 3 years at 40°C is shown in Fig. 6.4-4. It can be shown in the figure that a large portion of energy is wasted on the self-discharge of battery, 20%, the leakage of the supercapacitor, 11%, and the shutdown of all the components in the proposed converter topology, 24% in total.

If we extend the wanted service further, the minimum duration between adjacent activations for the low noise amplifier should be forced to be longer due to more energy consumption on the losses. The relationship of the calculated minimum duration between adjacent activations for a low noise amplifier over the wanted service life in years is shown in Fig. 6.4-5.

The minimum duration between adjacent activations for a low noise amplifier is 1.8 minutes for 3 years of operation, 3.3 minutes for 4 years of operation, 7.5 minutes for 5 years of operation, and 46 minutes for 6 years of operation. When the wanted service life is over 7 years, the total losses will surpass the total energy budget. Under the constraint of current energy budget, the proposed converter topology will operate for 6 years with the minimum duration between adjacent activations for a low noise amplifier to be 46 minutes.
The lithium manganese dioxide battery has the lower self-discharge than the silver oxide battery and seems to be the better choice when a longer service life is required. The problem with the lithium manganese dioxide battery is its drooping discharge profile. A single lithium manganese dioxide cell with a cell voltage of 3V can power the low noise amplifier and drive the LM2623 boost converter to supply a 100 \( \mu \)S pulse of 1A at 5V from the beginning. However, as the voltage going down the cell won't be able to drive the LM2623 boost converter to supply a
100 μS pulse of 1A at 5V and power the low noise amplifier. If there is a boost converter that can supply a 100 μS pulse of 1A at 5V with a minimum input voltage of 2V, the cut-off voltage of a single lithium manganese dioxide battery, we can then use the converter topology as shown in Fig. 6.4-6.

![Converter Topology Diagram](image)

Fig. 6.4-6 The converter topology using one lithium manganese dioxide battery as the input power source provided that a boost converter capable of supplying a 100 μS pulse of 1A at 5V with a minimum input voltage of 2V, the cut-off voltage of a single lithium manganese dioxide battery, can be found. A balancing circuitry is necessary for the supercapacitor.

The converter topology in Fig. 6.4-6 uses one lithium manganese dioxide battery as the input power source. Unlike the converter topology of Fig. 6.3-1, a balancing circuitry is necessary for the supercapacitor in this converter topology. The balancing circuitry will consume energy no matter what balancing approach is adopted. The balancing circuitry in Fig. 6.4-6 adopts the active approach, wherein a voltage comparator IC is included to consume 1.5 μA of supply current and the total leakage current would then add up to 2 μA in total.
Vout1 is 5V output for driving a power amplifier. When the cell voltage drops down to 2.75V, the minimum operating voltage for a low noise amplifier, a second boost stage becomes necessary. Vout2 is 3V output for driving a low noise amplifier and can be derived from Vout1, 5V, via a resistive divider. This converter topology still uses one boost converter to produce two different output voltages at the cost of more energy consumption from the longer activation of the boost converter to drive not only the power amplifier as in the previous converter topology, but also a low noise amplifier in this case.

We can follow the same procedure in this section to estimate the minimum duration between adjacent activations for a low noise amplifier based on the battery technology of a lithium manganese dioxide battery and the corresponding converter topology using one cell of lithium manganese dioxide battery as the input power source.

The comparison for the relationship of a calculated minimum duration between adjacent activations for a low noise amplifier over time between two cells of a silver oxide battery as the input power source and its corresponding converter topology in Fig. 6.3-1, and one cell of a lithium manganese dioxide battery as the input power source and its corresponding converter topology in Fig. 6.4-6 is shown in Fig. 6.4-7. The single cell of a lithium manganese dioxide battery as the input power source and its corresponding converter topology can manage to last longer for 6 years with the minimum duration between adjacent activations for a low noise amplifier to be 1 hour.
Fig. 6.4-7 The comparison for the relationship of the calculated minimum duration between adjacent activations for a low noise amplifier over time between two cells of silver oxide battery as the input power source and its corresponding converter topology in Fig. 6.3-1, and one cell of lithium manganese dioxide battery as the input power source and its corresponding converter topology in Fig. 6.4-6.

Even though the lithium manganese dioxide battery has a lower self-discharge rate than the silver oxide battery, the minimum duration between adjacent activations for a low noise amplifier for the converter topology using a silver oxide battery as shown in Fig. 6.3-1 is still shorter than that for the converter topology using a lithium manganese dioxide battery as shown in Fig. 6.4-6.
because the lower self-discharge rate of a lithium manganese dioxide battery is offset by the extra energy consumption necessary for the balancing circuitry in the converter topology.

In Fig. 6.4-5 and Fig. 6.4-7, the duration between adjacent activations $T$ is calculated based on the fixed pulse width and current level upon the activation of a specified low noise amplifier. The pulse width of a low noise amplifier will be changed if a different communication protocol is used and the current and voltage specification will also be changed if a different low noise amplifier is used in the converter topology in the future. How does the duration between adjacent activations $T$ change accordingly?

The product of pulse width $t$, current level $I$, and the operation voltage of the low noise amplifier is the energy consumption for an activation of the low noise amplifier, which is roughly proportional to the duration between adjacent activations $T$ since the total energy budget left for the activation of the low noise amplifier for a specified wanted service life is fixed.
6.5 Summary of Contribution

Silver oxide battery and lithium manganese dioxide battery were discharged at several constant current levels and at two different temperatures. The voltage and current levels were measured and the empirically determined constants of Peukert equation for each type of battery technology were curve-fitted. For each battery technology, it is shown from the results that there are two different sets of curve-fitting constants suited for discharge at small current levels and high current levels, respectively. The results for two types of battery technologies were compared. It is concluded from the experiments that the silver oxide battery is better than the lithium manganese dioxide battery in that the former has a less temperature-dependent battery constant, n, in the Peukert equation, a more flat discharge profile, and can deliver higher normalized power and energy than the latter does in spite of the fact that the latter has a higher energy density than the former. It is also found that the cut-off voltage for silver oxide battery can be considered as 1.5V at small discharge current levels instead of 1.2V as claimed by most manufacturers because the silver oxide battery would deliver 99% of nominal capacity before the voltage drops down to 1.5V when discharged at small current levels below 500 μA according to the measured result.

A converter topology using two silver oxide batteries in series producing an input voltage ranging from 3 to 3.1V based upon the unique characteristics of silver oxide battery from the aforementioned test result was proposed. There is only one boost converter in the proposed topology and no need for a balancing circuitry to balance the supercapacitor. Among all the converter topologies based upon the cell combinations of silver oxide battery and lithium manganese dioxide battery, the proposed converter topology is the simplest and the most energy-efficient.
7.1 Power Amplifier with Lower Supply Voltage

The converter topology will be the simplest if no voltage conversion is needed. To achieve this goal, a power amplifier with a lower supply voltage is necessary.

Fig. 7.1-1 The converter topology for a power amplifier with a supply voltage of 3V. The hybrid silver oxide battery/supercapacitor power source can drive the microcontroller, the low noise amplifier, and the power amplifier on the same voltage bus line without any voltage converters involved.

A shown in Fig. 7.1-1, the hybrid silver oxide battery/supercapacitor power source can drive the microcontroller, the low noise amplifier, and a power amplifier with a supply voltage of 3V on the same voltage bus line without any voltage converters involved. The power amplifier
with a lower supply voltage can demand more supply current to make its supply power equal to
the previous power amplifier with 5V input.

Furthermore, the topology in Fig. 7.1-1 can supply as much current as the power amplifier
demands as long as the voltage of the supercapacitor is above the minimum operating voltage of
the power amplifier and the current does not exceed the maximum current rating of the
supercapacitor, which is 30A for GW209D in this case.
7.2 Battery Technology with Lower Self-discharge

The self-discharge of battery cells consumes a greater portion of the energy budget as service time extends longer. When the demanded service life is over 6 years, neither a silver oxide battery nor a lithium manganese dioxide battery can retain enough capacity to perform normally for our application. The cause for high self-discharge of those battery technologies is due to the gradual drying-out of their fluid electrolyte. The battery with a solid electrolyte, such as lithium iodine batteries (Li/I\(_2\)), can perform extremely low self-discharge with a shelf life over 20 years. The power capability for the battery with a solid electrolyte is relatively low, usually in \(\mu W\) range, due to the high impedance of most solid electrolytes at normal ambient temperature. However, with the assistance of a supercapacitor most low drain batteries can fit into the aforementioned converter topologies and supply high power loads.

The discharge profile of lithium iodine batteries is similar to that of a lithium manganese dioxide battery and the nominal and cut-off voltages for a lithium iodine battery are 2.8V and 2V, respectively. The converter topology incorporating a single lithium iodine battery is similar to the one shown in Fig. 7.1. If a boost converter that can supply a 100 \(\mu S\) pulse of 1A at 5V with a minimum input voltage of 2V, the cut-off voltage of a single lithium iodine battery, can be found.
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APPENDIX A

EXPERIMENTAL SETUP

A.1 Introduction

One of the objectives of the experiment for this project is to test the candidate batteries under different discharge current levels and temperatures to obtain the discharge profiles of candidate batteries and thus predicting battery life.

The second objective is to test the leakage current of the supercapacitor under room and elevated temperatures and attain the leakage profile.

The third objective is to test the performance of the DC-DC converter with the help of the supercapacitor.

The experimental setup consists of five parts: the test circuitry, the oven if testing under an elevated temperature is required, the programmable DC power supply with a GPIB card, the data acquisition board, and the data acquisition software installed on a PC as shown in Fig. A.1-1.

The test circuitry is made to achieve the three aforementioned objectives and can be put in an oven where the temperature therein can be controlled. A programmable DC power supply is installed with a GPIB interface card to communicate with the PC, which makes the power supply programmable and configurable in several test modes through the data acquisition software. The data acquisition board acquires the voltage, current, and even the temperature data from the discharge circuitry via various types of transducers and then transfers the data to the data acquisition software.
Fig. A.1-1 Experimental setup system consisting of five parts: the test circuitry, the oven if testing under an elevated temperature is required, the programmable DC power supply with a GPIB card, the data acquisition board, and the data acquisition software installed on a PC.

The data acquisition software is the control center for the whole test. It sets the sampling rate of the incoming data, converts the analog incoming data into digital form, preprocesses the raw data, and saves them to files. It can also command the DC power supply via the GPIB interface into several test modes. When the data acquisition software detects that the incoming data has reached some pre-set conditions (say the battery voltage under certain level), it will shut
down the power supply and stop recording the data. The whole discharge test is implemented in automation and this so-called automatic test-bench can then be modified and expanded for future testing.

A.2 Test circuitry

The detailed schematic of the battery discharge circuitry is shown in Fig. A.2-1.

![Discharge Circuit Schematic](image)

Fig. A.2-1 Detailed schematic of the discharge circuit, where the DAQ is the short for the Data AcQuisition board.

The cell under test is held in a battery-holder for fixity and accessibility to the positive and negative terminals. A DC power supply is actually in series with the cell under test to force the cell to discharge. The insertion of a diode is to have a unidirectional current flow in the loop and prevent the cell from being charged by the power supply. The parameters being monitored and
recorded are the cell voltage, the discharge current, and even temperature in some cases. A 1Ω ± 5% resistor is used for current-sensing. Voltage drops across the current-sensing resistor and the terminals of the cell under test are sensed by the transducers of the data acquisition board (DAQ). The load resistance is inserted for controlling the discharge current level.

The detailed schematic of the leakage test circuitry for the supercapacitor is shown in Fig. A.2-2 (a) and (b). Any supercapacitors under test should be charged up by the power supply for at least three days before conducting the leakage test.

The supercapacitor under test is actually composed of two cells in series with a rated voltage 2.25V each to increase the voltage rating to 4.5V. When charging or discharging the supercapacitor, two balancing resistors of equal resistance should be in parallel with each cell to balance the voltage of each cell and prevent them from being over-charged. The resistance of the balancing resistors, as suggested by the manufacturer, can be any value from 39 KΩ to 100 KΩ. There are two ways of measuring leakage current. In Fig. A.2-2 (a), two cells in series are charged up at 4.5 Volts by a DC power supply and balanced by two balancing resistors for three days. The power supply and the balancing resistors are then removed and the voltage decay of the supercapacitor under test is recorded.
Fig. A.2-2 (a) and (b) Detailed schematics of the leakage test circuitry for the supercapacitor. (a) Two cells of the supercapacitor in series are charged up at 4.5 Volts by a DC power supply and balanced by two balancing resistors for three days before the test. The power supply and the balancing resistors are then removed and the voltage decay of the supercapacitor under test is recorded. The resistance of the balancing resistors, as suggested by the manufacturer, can be any value from 39 KΩ to 100 KΩ. (b) A single cell of the supercapacitor is charged up to a voltage within its rated voltage, 2.25 Volts via a charging/sensing resistor for three days. The voltage drop across the charging/sensing resistor is then recorded and converted to the equivalent leakage current. For a leakage current at the micro-amp level, the resistance of the charging/sensing resistor is determined to be 10 KΩ to be immune to measurement noises.

In A.2-2 (b), a single cell of the supercapacitor is charged up to a voltage within its rated voltage, 2.25 Volts via a charging/sensing resistor for three days. The voltage drop across the
charging/sensing resistor is then recorded and converted to the equivalent leakage current. For a 
leakage current at the micro-amp level, the resistance of charging/sensing resistor is determined 
to be $10 \text{ K}\Omega$ to be immune to measurement noises.

The test circuitry is also put into an oven for the leakage measurement at an elevated 
temperature.

The detailed schematic of the performance test circuitry for the DC-DC boost converter 
with the supercapacitor is shown in Fig. A.2-3.

![Fig. A.2-3 The detailed schematic of the performance test circuitry for the DC-DC boost converter with the supercapacitor.](image)

The output of the converter is connected to a programmable electronic load operating under the 
dynamic constant-current mode with a pre-settable current level and pulse duration. The power source charging up 
the supercapacitor could be either battery cells or a DC power supply. In both cases, a current-limiting resistor is 
added between the power source and the supercapacitor to mimic the nature of the limited current drain of the 
batteries. Two current sensing resistors of small resistance, 0.01$\Omega$ or 0.05$\Omega$, are inserted to sense the input and output 
currents, $i_{in}$ and $i_{out}$. 

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The output of the converter is connected to a programmable electronic load operating under the dynamic constant-current mode. The duration and frequency of the load current pulses can be programmed via the keypad at the front panel of the electronic load. The power source charging up the supercapacitor could be either battery cells or a DC power supply. In both cases, a current-limiting resistor is added between the power source and the supercapacitor to mimic the nature of the limited current drain of the batteries. Two current sensing resistors of small resistance, 0.01Ω or 0.05Ω, are inserted to sense the input and output currents. When the input voltage is below the rated voltage, 2.25 Volts, of a single cell of the supercapacitor in the case that one battery cell is used, only one cell of the supercapacitor is connected and no balancing resistors are needed. The load current, \( I_{\text{out}} \), is the controlled parameter with a pre-settable current level and pulse duration. The voltage of the supercapacitor, \( V_{\text{in}} \), the input current, \( I_{\text{in}} \), the output voltage, \( V_{\text{out}} \) and \( I_{\text{out}} \) are then recorded.

### A.3 Oven

One of the important issues in this project is the influence of temperature on the performance of the battery and the leakage current of the supercapacitor. For the test under a temperature other than room temperature, the test circuit is put into an oven, where the temperature there-within can be fully controlled. A thermometer is used for monitoring the actual temperature inside the oven to make sure that it meets the temperature setting of the oven. The test can proceed only when the temperature inside the oven stabilizes, usually thirty minutes after the temperature has been set.
A.4 Programmable DC Power Supply

The power supply for the battery discharge test is Xantrex XT 20-3 60Watts Programmable DC Power Supply [59]. The programmability of the power supply is achieved by the communication between the GPIB interface card installed at the rear panel of the power supply and the data acquisition software in the PC. The supply voltage and current level can be set manually by turning the round knobs on the front of the panel, or remotely using the data acquisition software from the PC. The power supply has two basic operating modes: Constant Voltage Mode and Constant Current Mode.

A.5 GPIB

GPIB is the short for General Purpose Interface Bus developed by Hewlett Packard in the late 1960’s and became the IEEE standard in 1975 also known as the IEEE 488 standard. A GPIB board installed on a PC is used to control and communicate with one or more external instruments that have a GPIB interface card, such as the aforementioned programmable DC power supply. GPIB was also updated and standardized by the IEEE, and was duly named IEEE 488.2. Some features of the GPIB are:

- It transfers data in parallel, one byte (eight bits) at a time.
- The hardware takes care of handshaking, timing, etc.
- Several instruments (up to 15) can be strung together on one bus.
- Data transfer is fast: 800 Kbytes/second or more.

The GPIB interface allows complete remote programming of the power supply, including status reporting, settings query, and interrupt generation with user-designated fault conditions.
Both the voltage and current output are precisely programmed directly in volts and amps with 16-bit resolution. The available programmable functions and specifications, detailed configuration and operation, and GPIB command format for XT 60 W series supply with a GPIB interface installed are listed in [60].

A.6 Data Acquisition System

A.6.1 Introduction

Data acquisition, or DAQ, is simply the process of measuring a real-world signal, such as a voltage, and bringing that information into the computer for processing, analysis, storage, or other data manipulation. Obtaining proper results from a PC-based DAQ system depends on each of the following system elements (see Fig. A.6.1-1):

- The PC
- Transducers
- Signal Conditioning
- DAQ Hardware
- Software
A.6.2 Data Acquisition Hardware

The data acquisition hardware used in this project is National Instruments Corporation’s 6023E [62]. It features 16 channels of analog input, two 24-bit counters, a 68-pin connector and eight lines of digital I/O. It has a 12-bits resolution and a sampling rate up to 200k per second.

A.6.3 Data Acquisition Software

LabVIEW ([61]), short for Laboratory Virtual Instrument Engineering Workbench, was first developed by National Instruments in 1983 in an effort to minimize the time needed to program instrumentation systems. It is a programming environment in which one creates programs with graphics; in this regard it differs from traditional programming language like C, C++, or Java, in which one programs with text. Whereas other programming systems use text-based language to create lines of code, LabVIEW uses a graphical programming language to create programs in a pictorial form called a block diagram, thus eliminating a lot of the syntactical details.

LabVIEW uses terminology, icons, and ideas familiar to scientists and engineers. It relies
on graphical symbols rather than textual language to describe programming actions. The principle of dataflow, in which functions execute only after receiving the necessary data, governs execution in a straightforward manner.

LabVIEW programs are called virtual instruments (VIs) because their appearance and operation imitate actual instruments. However, behind the scenes they are analogous to main programs, functions, and subroutines from popular programming language like C or BASIC. A VI has three main parts:

- The **Front Panel** is the interactive user interface of a VI, named because it simulates the front panel of a physical instrument. The front panel can contain knobs, push buttons, graphs, and many other controls (which are user inputs) and indicators (which are program outputs). A user will input data using a mouse and keyboard and then view the results produced by the program on the screen.

- The **Block Diagram** is the VI's source code, constructed in LabVIEW's graphical programming language, G. The block diagram is the actual executable program. The components of a block diagram are lower level VIs, built-in functions, constants, and program execution control structures. One draws wires to connect the appropriate objects together to indicate the flow of data between them. Front panel objects have corresponding terminals on the block diagram so that data can pass from the user to the program and back to the user.

- In order to use a VI as a subroutine in the block diagram of another VI, it must have an **icon** and a **connector**. A VI that is used within another VI is called a *subVI* and is
analogous to a subroutine. The icon is a VI’s pictorial representation and is used as an object in the block diagram of another VI. A VI’s connector is the mechanism used to wire data into the VI from other block diagrams when the VI is used as a subVI. Much like parameters of a subroutine, the connector defines the inputs and outputs of the VI.

VIs are hierarchical and modular. One can use them as top-level programs or sub-programs. With this architecture, LabVIEW promotes the concept of modular programming. First, one divides an application into a series of sub-tasks. Next, one builds a VI to accomplish each sub-task and then combines those VIs on a top-level block diagram to complete the larger task.

Between NI-DAQ hardware and LabVIEW software, there is a utility called MAX (Measurement and Automation Explorer). MAX is a Windows software interface that gives one access to all National Instruments boards, one of which is 6023E in this experiment. MAX is mainly used for configuring and testing the hardware. This is very useful to do before one attempts to access the hardware in LabVIEW. MAX is installed by default when one installs LabVIEW. Fig. A.6-2 shows the relationships among LabVIEW, MAX, and NI-DAQ.
A DAQ board has several analog I/O parameters that control the operation of the ADC (analog-to-digital converter) and DAC (digital-to-analog converter). On almost all boards, one will configure these settings in the MAX software utility. The types of settings on 6023E DAQ boards include the following:

- **ADC Input Range**: Unipolar 0 to +10 V
  - Bipolar ±5 V
  - Bipolar ±10 V (default)

- **ADC Input Mode**: Ground-referenced single-ended
  - Non-referenced single-ended
  - Differential (default)

- **DAC Reference**: Internal (default)
  - External
- DAC Polarity: Unipolar—straight binary mode
  Bipolar—Two’s complement mode (default)

The functionality of MAX is divided into four categories:

- **Data Neighborhood:** It shows all of the currently configured virtual channels and provides utilities for testing and reconfiguring those virtual channels and creating new virtual channels. A virtual channel is a shortcut to a configured channel in the data acquisition system. One can give the channel a description, decide what type of transducer the channel will use, set the range, choose the grounding mode, assign custom scaling for your virtual channel, and give the channel a descriptive name to replace the channel number all at the same time. In this experiment, the signals to be measured are the battery voltage and discharge current and hence two virtual channels are needed. Since the differential-measurement mode is chosen for the experiment, each virtual channel should be configured to include a pair of physical channels on the DAQ board, say ACH0 and ACH1, for differential measurement.

- **Devices and Interfaces:** It shows any currently installed and detected National Instruments hardware, such as DAQ boards and GPIB boards. Devices and Interfaces also include utilities for configuring and testing devices. There are three options for the installed board or device: Properties, Test Panels, and Delete. The Properties panel is where one can actually configure the DAQ board, like 6023E in this experiment. Under the Properties panel, there is an AI tab. The AI tab configures the analog input signal, specifies the type of signal, its voltage limits, and what scale to map it to. In
this experiment, the analog input signals are the signals to be measured, the battery voltage and discharge current.

- **Scales:** Here one can create custom scales one can use to determine scaling information for existing virtual channels. This is sometimes necessary or useful particularly for sensors that are not linear or where one wants to read the actual units directly instead of having to convert voltage or current to the desired unit, such as temperature. In this experiment, there's no scaling for current measurement since the measured data is actually the voltage drop across a $1\,\Omega$ resistor. It is a one-to-one scaling according to Ohm’s law.

- **Software:** It shows all of the currently installed versions of National Instruments software.

### A.6.4 Virtual Instrument (VI) for Battery Discharge Test

There are several tasks to be done by software for the battery discharge test:

- **Task 0** Configuring virtual channels and their corresponding settings.
- **Task 1** Specifying the file path and name for storing the data
- **Task 2** Specifying the required voltage or current level for the programmable DC power supply via GPIB interface.
- **Task 3** Specifying the input channels, the sampling rate, and number of total scans for one data point.
Task 4  Specifying the duration between adjacent data acquisitions, and the end (cutoff) voltage for the battery under testing that is considered as the end of the experiment.

Task 5  Acquiring the data according to the sampling rate and number of total scans for one data point as specified in Task 3, averaging the acquired data according to the number of total scans, and storing the averaged data point to the file as specified in Task 1. The subsequent data points will be appended to the same file in order. Waiting for the duration as specified in Task 4 and then performing the next acquisition.

Task 6  Terminating the experiment when the averaged data point representative of the battery voltage drops below to the end (cutoff) voltage as specified in Task 4, and resetting the programmable DC power supply.

Task 0 has to be done in MAX before the discharge test, while all other tasks are implemented in LabVIEW. Fig. A.6-3 shows the data flow diagram of Virtual Instrument (VI) for the Battery Discharge Test.
Fig. A.6-3 The data flow diagram of Virtual Instrument (VI) for the battery discharge test. There are three subVIs, GPIB.vi, Al.vi, and Write to Spreadsheet.vi. The pins with number beside them are the user-specified inputs to the corresponding subVIs.
There are three subVIs, GPIB.vi, AI Acquire Waveforms.vi, and Write to Spreadsheet.vi. The pins with number beside them are the inputs to the corresponding subVIs.

a) AI Acquire Waveforms

The functional schematic for the subVI, **AI Acquire Waveforms.vi**, is shown in Fig. A.6-4.

![functional schematic for AI Acquire Waveforms.vi](image)

Fig. A.6-4 The functional schematic for the **AI Acquire Waveforms.vi** ([61])

This VI acquires the data from the specified virtual channels on the specified device, if more than one DAQ boards are available, and samples the virtual channels for the specified number of samples per channel at the specified sampling rate. The output of the VI is a 1-D data array or 2-D data array if data from multiple virtual channels is acquired at the same time. In this experiment, only one DAQ board is available, NI-6023E. There are two virtual channels for acquisition, one for voltage measurement and the other for current measurement. The number of samples per channel is set to be 100, so the output of the VI is a 2x100 data array. Although the maximum scan rate for NI-6023E DAQ board is 200 kS/s, the scan rate for multiple channel acquisitions should be set to a lower value for greater data precision. Hence, the scan rate is set to
be 80 kS/s in this experiment.

We only need the average readings of each acquisition for each channel, so each 100 samples from each channel are averaged to get one data point. Now the previous 2x100 data array is converted into 2x1 data array and will be sent to the next subVI for storage.

Since some of the tests could last for several days if discharged by a very small current, we cannot acquire data continuously and hence there is a reasonable delay necessary for acquiring the next data points. This delay or duration between acquisitions, can be specified by the user in LabVIEW.

The averaged data from the voltage measurement channel is examined after each acquisition to see if it drops down to the specified end voltage value, meaning the test will stop if the battery voltage drops down to the specified level.

b) Write to Spreadsheet File

The functional schematic for the subVI, Write To Spreadsheet File.vi, is shown in Fig. A.6-5. This VI converts a 1-D or 2-D array of single-precision numbers to a text string and writes the string to a new byte stream file or appends the string to an existing file as specified by the user. The data to be stored will be sent to the “1D data” input port if only one channel of data is acquired or “2D data” input port if multiple channels of data are required as shown in Fig. A. 6-5.

In this experiment, the Write to Spreadsheet File.vi receives the averaged data from AI Acquire Waveforms.vi and saves it to a spreadsheet file, say Excel, as specified in the “file path” input port as shown in Fig. A.6-5. The consequent acquired data will be appended to the same spreadsheet file in order.
c) GPIB.vi

The functional schematic for the subVI, GPIB.vi, is shown in Fig. A.6-6.

A device connected to a GPIB bus can be written to or be read from using this VI. Select the GPIB address, choose Read or Write or both, type in the characters to be written, and run the VI.

The programmable DC power supply has to be set in the Remote Control Mode before we
can remotely control it by software. In “Characters to write” input port we can write the GPIB command, VSET 20V, to set the voltage of the power supply to be 20 Volts. When the whole experiment stops, the command “CLR” is issued to the power supply to CLEAR any current setting and return to the power-on state.
A Stellar sealion can take an underwater dive as deep as 500m according to zoological survey literatures, so it is important to know whether or not the components involved in the tag can sustain such a high pressure. Table B-1 lists the gravimetric and volumetric rates of change after being pressurized with predetermined pressures for the Maxell SR1130 silver oxide battery, the Panasonic CR2412 lithium manganese dioxide battery, and cap-XX GW209D supercapacitor.

Table B-1 Volumetric rate of change in percentage after being pressurized with predetermined pressures

<table>
<thead>
<tr>
<th>Depth</th>
<th>Maxell SR1130W</th>
<th>Panasonic CR2412</th>
<th>GW209D Supercapacitor</th>
</tr>
</thead>
<tbody>
<tr>
<td>125psi/80m depth</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>225psi/152.6m depth</td>
<td>~0</td>
<td>~0</td>
<td>~0</td>
</tr>
<tr>
<td>325psi/225.2m depth</td>
<td>~0</td>
<td>~0</td>
<td>~0</td>
</tr>
</tbody>
</table>

All the components under pressurized test were pressurized for at least 16 hours, and their weights and sizes were recorded before and after a test. The test pressure starts from 125psi and has an incremental pressure of 100psi.

It showed that all the components under test could hold up to 325psi, equivalent to the pressure at about 225.2m underwater, without substantial physical deformation. However, one of supercapacitors under test has cracked open at 225psi. It is necessary to provide supercapacitors with solid casing to resist underwater pressure.