Copyright Statement

The digital copy of this thesis is protected by the Copyright Act 1994 (New Zealand).

This thesis may be consulted by you, provided you comply with the provisions of the Act and the following conditions of use:

- Any use you make of these documents or images must be for research or private study purposes only, and you may not make them available to any other person.
- Authors control the copyright of their thesis. You will recognize the author's right to be identified as the author of this thesis, and due acknowledgement will be made to the author where appropriate.
- You will obtain the author's permission before publishing any material from their thesis.

General copyright and disclaimer

In addition to the above conditions, authors give their consent for the digital copy of their work to be used subject to the conditions specified on the Library Thesis Consent Form and Deposit Licence.
Wireless Power Enabled
Ultra-Energy Efficient
Bi-Stable Actuator for Microfluidics

Dulsha Kularatna-Abeywardana

A thesis submitted in partial fulfilment of the requirements for the degree of
Doctor of Philosophy in Electrical and Electronic Engineering,
The University of Auckland, New Zealand.

July 2018
Abstract

In the past few decades microfluidic systems have gained significant interest both in academia and industry. They are used for chemical analysis where small quantities of fluids are treated using special components such as microvalves, micropumps and micromixers. The automated biochemical laboratory is a microfluidic platform that studies the toxic effects of chemicals on Zebra fish embryos. This platform operates several solenoid microvalves for very long hours leading to a large power consumption. To improve the portability of the system the power consumption needs to minimised to reduce the required battery capacity. This thesis aims to develop a new energy efficient microactuator to reduce the power consumption of these solenoid microvalves.

The thesis proposes a novel microfluidic actuator for microvalves, with bi-stable actuation, wireless power and wireless actuation control, which is significantly more efficient than state of the art microvalves reported in literature. An ultra-energy efficient bi-stable actuation mechanism, primary and secondary wireless power supply circuits and wireless actuation control has been designed and implemented. The actuation mechanism, magnetic and electrical circuits of the actuator and power supply have been theoretically modelled and analysed. Here the actuator is driven by electropermanent magnets, together with an inductively coupled wireless power transfer system and supercapacitor based energy buffering. Energy gradually extracted through the wireless power transfer system, accumulated on the supercapacitor buffer, provides momentary peak power for actuation only. The microactuator also features a unique communication mechanism to deliver the valve control signal to the microactuator.

It has been shown experimentally that the new microactuator reduces the energy consumption of the automated biochemical laboratory by an order of two. This microactuator also features ultra-fast, bi-stable actuation, significant deflection and force capabilities. This novel microactuator provides wireless power and control, ultra-energy efficient actuation, high deflection and force capability, making them a unique class of microfluidic actuators.
To my loving family Rajith, Nethuli and Mineli......
Acknowledgements

First, I would like to thank the Faculty of Engineering at the University of Auckland, for offering me the Faculty of Engineering Doctoral Scholarship, which made a big difference in facing this tedious and testing journey of studying towards a PhD.

I would like to thank my supervisor Dr. Patrick Hu for his invaluable support, guidance and motivation throughout my PhD. His expertise in the area of power electronics has been a major advantage for my research. He has supported me in many ways to get through this PhD, encouraging me at difficult times.

I would also like to extend my gratitude to my second supervisor Professor Zoran Salcic, for his guidance, support and encouragement towards completing this research.

I would also like to thank Dr. Kevin Wang, who has been there throughout this research, guiding me and supporting me in achieving my goals. He has been a very motivating and supportive of my research and this has meant a lot to me.

I would like to extend my gratitude to other postgraduate students, Dr. Wei Chen, Mr. Kun He and Mr. Mozammil Ahsan who worked alongside with me and assisted me in completing various parts of this research.

In addition, I would like to thank the technical staff in the ECE department, especially Mr. Howard Lu, Mr. Rob Champion, Mrs. Kavitha Penneru, Mr. Akshat Bisht and Ms. Weiwei Ai for their technical support and laboratory assistance. I am also very grateful to the technical staff in the Department of Chemical and Materials Engineering for their assistance in some of my experimental work.

I would also like to thank my friends at the electrical department Dr. Jianlong Tian, Dr. Liang Huang, Dr. Saranga Weerasinghe, Dr. Adeel Zaheer, Ms. Maryam Hemmati, Mr. Hamidreza Rahanmaee, Ms. Yuan Liu, Mr. Yuan Song, Mr. Rong Hua, Mr. Bui Van Dai,
Ms. Hanie Mehdinezhad, Ms. Hoda Rezaie, Ms. Latha Murugesan, Mr. Andy Leung, Ms. Suchetha Sehgal, Ms. Jackie Zou, Ms. Thiaza Thasthakeer and other peers for their guidance, support and companionship during this journey.

This achievement would not have been possible without the support from my extended network of friends outside the university who helped us with babysitting, meals and moral support at difficult times. Thank you Chanika, Nadun, Hasini, Dhanu, Keshav, Cecilia, Aunty Manel, Uncle Dayajith, Aunty Jayantha, Uncle Dayananda, Aunty Upendra, Uncle Kumar, and all others who helped me along the way in numerous ways.

I would like to extend my gratitude to our families, my parents Nihal and Priyani, sister Malsha and her husband Kasun, husband’s parents Abey and Ranjini, aunt Ramani and sister in law Ruwandhi for all their support. I would especially like to mention my mum’s dedication towards keeping my fridge loaded with cooked food every week, delivered straight from her kitchen in Hamilton. My dad has always been my inspiration, and told me repeatedly to be a ‘real engineer’ and chants ‘burning is learning!’ His critical comments on my work has always helped me produce meaningful results.

Thank you to my little daughters Nethuli and Mineli, who do not quite understand the concept of a PhD, but have been the ultimate joy at the end of a long day of intensive research. I hope, that despite my absence and having to put your needs on the side-lines at times, I have inspired you to dream big and work hard to make your dreams a reality.

Last but not least, my heartfelt gratitude goes out to my loving husband Rajith, who not only supported me at each step, in every possible way to achieve this milestone in my life but also decided to do a PhD along with me! You made the late nights, countless disappointments, parenting and doctoral research a lot more bearable. You have always inspired me to dream big and helped me make my dreams come true.

This thesis would not have been possible without each and every one of you, so thank you!

Dulsha Kularatna-Abeywardana

21st September 2017
# Table of Contents

ABSTRACT ...................................................................................................................... i

ACKNOWLEDGEMENTS .................................................................................................... iii

LIST OF FIGURES ............................................................................................................. ix

LIST OF TABLES ............................................................................................................... xiii

NOMENCLATURE ............................................................................................................ xiv

Chapter 1

INTRODUCTION ............................................................................................................... 1

1.1. Microfluidic Systems ................................................................................................. 1

1.1.1. Microfluidic Applications .................................................................................... 3

1.1.2. Microfluidic Components .................................................................................... 5

1.2. Automated Bio-Chemical Laboratory ...................................................................... 6

1.2.1. Microfluidic Valves in ABL Platform .................................................................. 9

1.2.2. Requirement of a Novel Valve Actuator ............................................................. 10

1.3. Wireless Power Supplies ......................................................................................... 11

1.4. Objectives and Scope of Research .......................................................................... 13

1.5. Thesis Organisation .................................................................................................. 15

Chapter 2

LITERATURE REVIEW ................................................................................................... 17

2.1. Microvalves ............................................................................................................... 17

2.1.1. Performance Parameters of Microvalve Actuators ............................................. 20

2.1.2. Electromagnetic Microvalve Actuators ............................................................... 21

2.2. Solenoid Valves in ABL Platform .......................................................................... 23

2.2.1. Solenoid Valve Operation in Experimental Run ................................................ 24

2.2.2. Drawbacks of Solenoid Valves .......................................................................... 25

2.3. State-of-the-Art Microvalves ................................................................................... 26
Chapter 3

PROPOSED BI-STABLE MICROACTUATOR ................................................................. 30

3.1. Electromagnetic Actuation ........................................................................... 33
3.2. Wireless Power and Energy Buffer .............................................................. 33
3.3. Wireless Control of Actuator ....................................................................... 34

Chapter 4

BI-STABLE ACTUATION BASED ON ELECTRO-PERMENANT MAGNETS .......... 36

4.1. Electro-Permanent Magnet .......................................................................... 36
   4.1.1. Principles of Operation .......................................................................... 37
   4.1.2. Material Selection .................................................................................. 39
4.2. Electromagnetic Modelling and Analysis ..................................................... 41
   4.2.1. Magnetic Circuit Modelling .................................................................... 41
   4.2.2. Electric Circuit Modelling ................................................................. 44
   4.2.3. Magnetic Simulation Analysis .............................................................. 46
4.3. Actuation Control Design and Simulation ................................................... 48
   4.3.1. Magnetic Excitation Circuit Design ...................................................... 48
      4.3.1.1. Main Control Circuit ....................................................................... 49
      4.3.1.2. Actuator Position Detection ........................................................... 50
   4.3.2. Actuator Control Circuit Simulation ...................................................... 53
4.4. Actuator Implementation ............................................................................. 54
4.5. Experimental Results and Analysis ............................................................. 57
   4.5.1. Number of Turns versus Energy Consumption ..................................... 57
   4.5.2. Pulse Duration versus Energy Consumption ......................................... 60
   4.5.3. Holding and Attraction Force of Actuator .............................................. 62
   4.5.4. Deflection Range of Actuator ............................................................... 65
4.6. Performance Summary .............................................................................. 67

Chapter 5

WIRELESS POWER SUPPLY WITH SUPERCAPACITOR ENERGY BUFFER ........ 69

5.1. Review of Wireless Power Supply Technologies ......................................... 69
<table>
<thead>
<tr>
<th>Section</th>
<th>Title</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>5.2.</td>
<td>Inductive Power Transfer for Microactuator</td>
<td>72</td>
</tr>
<tr>
<td>5.3.</td>
<td>Wireless IPT Power Supply Design</td>
<td>75</td>
</tr>
<tr>
<td>5.3.1.</td>
<td>IPT Setup</td>
<td>76</td>
</tr>
<tr>
<td>5.3.2.</td>
<td>Track and Coil Design</td>
<td>77</td>
</tr>
<tr>
<td>5.3.3.</td>
<td>Tuning Circuit Design</td>
<td>80</td>
</tr>
<tr>
<td>5.3.2.1.</td>
<td>Operating Frequency Selection</td>
<td>81</td>
</tr>
<tr>
<td>5.3.2.2.</td>
<td>Compensation Network</td>
<td>83</td>
</tr>
<tr>
<td>5.3.2.3.</td>
<td>Quality Factor</td>
<td>85</td>
</tr>
<tr>
<td>5.3.4.</td>
<td>Primary Converter Design</td>
<td>86</td>
</tr>
<tr>
<td>5.3.3.1.</td>
<td>Primary Converter Topologies</td>
<td>86</td>
</tr>
<tr>
<td>5.3.3.2.</td>
<td>Analysis of Current-fed Autonomous Push-Pull Converter</td>
<td>88</td>
</tr>
<tr>
<td>5.3.3.3.</td>
<td>Voltage-fed Full Bridge Converter</td>
<td>98</td>
</tr>
<tr>
<td>5.3.5.</td>
<td>Secondary Pick-Up Circuit Design</td>
<td>102</td>
</tr>
<tr>
<td>5.4.</td>
<td>Energy Buffer Design</td>
<td>104</td>
</tr>
<tr>
<td>5.4.1.</td>
<td>Overview of Energy Storage Technologies</td>
<td>105</td>
</tr>
<tr>
<td>5.4.2.</td>
<td>Supercapacitor Technology</td>
<td>108</td>
</tr>
<tr>
<td>5.4.3.</td>
<td>Supercapacitor Energy Buffer</td>
<td>110</td>
</tr>
<tr>
<td>5.3.3.4.</td>
<td>Theoretical Modelling of Supercapacitor Energy Buffer</td>
<td>110</td>
</tr>
<tr>
<td>5.3.3.5.</td>
<td>Electrical Circuit Design of Energy Buffer</td>
<td>113</td>
</tr>
<tr>
<td>5.3.3.6.</td>
<td>Simulation Study of Supercapacitor Energy Buffer</td>
<td>114</td>
</tr>
<tr>
<td>5.5.</td>
<td>Wireless Power Platform Development</td>
<td>115</td>
</tr>
<tr>
<td>5.6.</td>
<td>Experimental Results</td>
<td>118</td>
</tr>
<tr>
<td>5.6.1.</td>
<td>Push Pull Converter Performance</td>
<td>118</td>
</tr>
<tr>
<td>5.6.2.</td>
<td>Full Bridge Converter Performance</td>
<td>120</td>
</tr>
<tr>
<td>5.6.3.</td>
<td>Comparison of Push Pull and Full Bridge Converters</td>
<td>122</td>
</tr>
<tr>
<td>5.6.4.</td>
<td>Pick Up Circuit Performance</td>
<td>123</td>
</tr>
<tr>
<td>5.6.5.</td>
<td>Supercapacitor Energy Buffer Performance</td>
<td>125</td>
</tr>
<tr>
<td>5.7.</td>
<td>Performance Summary</td>
<td>125</td>
</tr>
</tbody>
</table>
Chapter 6

Wireless Actuation Control Based on Frequency Variation ............... 127

6.1. Short Range Wireless Communication Technologies .................. 128

6.2. IPT Based Wireless Communication ............................................. 130

6.2.1. Frequency Detection Techniques ............................................ 131

6.2.1.1. Microcontroller Driven Frequency Counter ......................... 132

6.2.1.2. Frequency to Voltage Converter ........................................ 133

6.2.1.3. Analog Passive Bandpass Filter ........................................ 134

6.2.2. Control Signal Detector Design ............................................. 135

6.2.2.1. Filter Type Selection ...................................................... 135

6.2.2.2. Actuator Control Circuit Design ....................................... 136

6.2.2.3. Wireless Control Implementation ...................................... 137

6.2.3. Performance Analysis of Communication Module ................... 138

Chapter 7

Overall Performance of Bi-Stable Wireless Microactuator ............... 141

7.1. Full System Operation .............................................................. 141

7.2. Summary of Microactuator Performance ................................... 142

7.3. Performance Comparison with Existing Microvalve Actuators ....... 144

7.3.1. Comparison with Normally Open Solenoid Pinch Valve ............ 144

7.3.2. Comparison with Other Active Microvalves ............................ 146

7.3.3. Comparison with Bi-Stable Microvalves ................................. 147

Chapter 8

Conclusions and Future Recommendations ...................................... 149

8.1. General Conclusions ............................................................... 149

8.2. Thesis Contributions ............................................................... 154

8.3. Publications ............................................................................. 155

8.4. Recommendations for Future Work ........................................... 156

References ....................................................................................... 159
List of Figures

Figure 1.1: Channel based microfluidic systems [6-8] ................................................................. 2
Figure 1.2: Droplet based microfluidic systems [6, 9, 10] ............................................................. 2
Figure 1.3: Microfluidic Systems [6, 11, 12] ............................................................................. 3
Figure 1.4: Real-life applications of microfluidic systems [17-20] ............................................. 4
Figure 1.5: Microfluidic flow control components [23-26, 29, 31] ............................................ 6
Figure 1.6: Automated Bio-chemical Laboratory (ABL) [21] .................................................... 7
Figure 1.7: Photographs of ABL platform .................................................................................. 8
Figure 1.8: Commercially available wireless power devices [45, 46] ....................................... 12
Figure 2.1: Classification of microvalves ..................................................................................... 18
Figure 2.2: Externally actuated active microvalve mechanisms [57] ........................................... 19
Figure 2.3: Energy density range of microvalve actuators [51] .................................................. 20
Figure 2.4: Electromagnetic actuator configurations [51] ......................................................... 22
Figure 2.5: BioChem™ pinch valve and its operational principle [24] .......................................... 24
Figure 3.1: Features of proposed microvalve actuator ................................................................. 30
Figure 3.2: Microvalve actuator overview .................................................................................... 32
Figure 3.3: Interaction between modules in proposed microactuator and WPT system ............. 32
Figure 3.4: Basic concept of electropermanent magnet actuation ............................................. 33
Figure 3.5: Basic wireless power supply layout ......................................................................... 34
Figure 3.6: Logical flow of the microactuator control ................................................................. 35
Figure 4.1: EP magnet actuation concept .................................................................................... 36
Figure 4.2: Magnetisation curves for soft, semi-hard and hard magnetic materials .................. 39
Figure 4.3: Magnetisation curves for AlNiCo and NdFeB ............................................................ 41
Figure 4.4: Equivalent magnetic circuit for EP magnet actuator ............................................... 42
Figure 4.5: Dimensions of coil and magnetic structure ............................................................... 43
Figure 4.6: Equivalent circuit for magnetizing coil ................................................................. 45
Figure 4.7: 3D view of FEMM simulation of external magnetic field generation ................. 47
Figure 4.8: Top view of FEMM simulation ............................................................................... 47
Figure 4.9: Cross section view of AlNiCo during FEMM simulation ..................................... 48
Figure 4.10: Actuator control circuit ....................................................................................... 49
Figure 5. 15: Resonant tank and gate voltages .................................................................................. 94
Figure 5. 16: Steady state analysis for ZVS switching frequency ......................................................... 94
Figure 5. 17: Simulation results of current fed push-pull converter .................................................... 98
Figure 5. 18: Voltage-fed full bridge primary converter ...................................................................... 99
Figure 5. 19: Fixed-frequency switching of voltage-fed full bridge converter .................................... 100
Figure 5. 20: NI9401 Module connected to the CompactRio PID controller ....................................... 101
Figure 5. 21: LT spice simulation results for voltage fed full bridge converter ................................. 102
Figure 5. 22: Secondary pick up circuit .............................................................................................. 103
Figure 5. 23: LT Spice simulation results for voltage fed full bridge converter ......................... 104
Figure 5. 24: Rechargeable energy storage technologies ................................................................. 105
Figure 5. 25: Specific power against specific energy for rechargeable energy storage options [104] ......................................................................................................................... 107
Figure 5. 26: Charge-Discharge curve of battery and supercapacitor [107] ........................................... 108
Figure 5. 27: Internal electrode arrangement of supercapacitors [110] ................................................ 109
Figure 5. 28: Supercapacitors available in the market today [104, 113-115] ...................................... 110
Figure 5. 29: Electric circuit model of supercapacitor ...................................................................... 110
Figure 5. 30: Internal resistance against depth of discharge of a supercapacitor and battery [116] ....................................................................................................................................... 112
Figure 5. 31: Series connected supercapacitor energy buffer .............................................................. 113
Figure 5. 32: Convertible series-parallel arrangement of supercapacitors ........................................ 114
Figure 5. 33: Charging cycle of supercapacitor energy buffer ........................................................... 114
Figure 5. 34: Track and coil setup of wireless power supply ............................................................... 115
Figure 5. 35: Initial implementation of autonomous push-pull converter ........................................ 116
Figure 5. 36: H-bridge primary converter ......................................................................................... 116
Figure 5. 37: Secondary pick-up circuit ............................................................................................ 117
Figure 5. 38: Supercapacitor energy buffer ....................................................................................... 117
Figure 5. 39: Resonant and non-resonant operation of autonomous push pull converter ................. 119
Figure 5. 40: Pick up power against primary track current ................................................................. 120
Figure 5. 41: Gate drive signal for full bridge converter at 100 kHz, 800 kHz and 1 MHz ............. 121
Figure 5. 42: Output of the primary converter and pick up circuit at 800 kHz ............................... 121
Figure 5. 43: Power delivered to microactuator against primary track current ............................... 122
Figure 5. 44: Output power against primary track current ............................................................... 122
Figure 5. 45: Rectified output voltages for different pick up voltages ............................................. 124
Figure 5. 46: Supercapacitor charging curve ..................................................................................... 125
Figure 6.1: Basic block diagram of control signal communication........................................... 131
Figure 6.2: Microcontroller driven frequency detector .......................................................... 132
Figure 6.3: Frequency to voltage converter circuit............................................................... 133
Figure 6.4: Frequency response of the filter ......................................................................... 134
Figure 6.5: 3rd Order bandpass filter with 1 MHz centre frequency ...................................... 137
Figure 6.6: Frequency and phase response of the 3rd Order bandpass filter ......................... 137
Figure 6.7: Implemented control signal communication module .......................................... 138
Figure 6.8: Frequency response of bandpass filter with centre frequency of 1 MHz .......... 139
Figure 6.9: Filter output at IPT power transfer frequency ..................................................... 139
Figure 6.10: Filter output at IPT power transfer frequency of 800 kHz ............................. 140
Figure 7.1: Valve control process for opening valve .............................................................. 142
Figure 7.2: Novel actuator performance comparison with solenoid valve actuator in ABL platform.................................................................................................................. 145
Figure 7.3: Power and energy performance of EP magnet actuator vs current solenoid valves in target microfluidic platform ....................................................................................... 145
Figure 7.4: Power and energy consumption of EP magnet actuated valve in comparison with other externally actuated microvalves ........................................................................... 147
List of Tables

Table 2.1: Power and energy consumption of microvalves in ABL platform ........................................... 24
Table 2.2: Performance of low power latching valves ................................................................. 29
Table 4.1: Coercivity and remanence of magnetic materials ......................................................... 40
Table 4.2: Tact switches investigated for position feedback circuit .............................................. 52
Table 4.3: Dimensions of magnetic structure ................................................................................... 54
Table 4.4: Summary of actuator performance ................................................................................... 67
Table 5.1: Comparison of main WPT systems [86-89] ................................................................. 71
Table 5.2: Output properties of different pick-up compensation circuits ........................................ 84
Table 5.3: Power converter topology comparison ................................................................. 87
Table 5.4: Circuit parameters used for LT Spice simulation ...................................................... 97
Table 5.5: Circuit parameters used for LT Spice simulation of voltage fed converter ................. 101
Table 5.6: Comparison of energy storage technologies [104] ..................................................... 106
Table 5.7: Performance summary of different pick up coils ....................................................... 118
Table 5.8: Pick-up inductor capacitor combinations experimented .............................................. 120
Table 5.9: Summary of wireless power supply performance .................................................... 126
Table 6.1: Comparison of short range wireless communication technologies [118-122] .......... 130
Table 6.2: Comparison of Butterworth, Chebyshev, Bessel and Elliptic filters [123, 124] ........... 135
Table 6.3: Chebyshev filter specifications for microactuator control .......................................... 136
Table 7.1: Operation parameters of novel microactuator .............................................................. 143
Table 7.2: Comparison of EP magnet actuator and state-of-the-art bistable valves .................... 148
# Nomenclature

## Acronyms

<table>
<thead>
<tr>
<th>Acronym</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>A4WP</td>
<td>Alliance for Wireless Power</td>
</tr>
<tr>
<td>ABL</td>
<td>Automated bio-chemical laboratory</td>
</tr>
<tr>
<td>AC</td>
<td>Alternating current</td>
</tr>
<tr>
<td>AFC</td>
<td>Alkaline fuel cell</td>
</tr>
<tr>
<td>BLE</td>
<td>Bluetooth low energy</td>
</tr>
<tr>
<td>CMOS</td>
<td>Complementary metal-oxide-semiconductor</td>
</tr>
<tr>
<td>CPT</td>
<td>Capacitive power transfer</td>
</tr>
<tr>
<td>DC</td>
<td>Direct current</td>
</tr>
<tr>
<td>DNA</td>
<td>Deoxyribonucleic acid</td>
</tr>
<tr>
<td>ELDC</td>
<td>Electric double layer capacitors</td>
</tr>
<tr>
<td>EMI</td>
<td>Electromagnetic interference</td>
</tr>
<tr>
<td>EP</td>
<td>Electropermanent</td>
</tr>
<tr>
<td>ESR</td>
<td>Equivalent series resistance</td>
</tr>
<tr>
<td>EV</td>
<td>Electric vehicle</td>
</tr>
<tr>
<td>FEMM</td>
<td>Finite element method magnetics</td>
</tr>
<tr>
<td>FPGA</td>
<td>Field programmable gate array</td>
</tr>
<tr>
<td>FVC</td>
<td>Frequency to voltage converter</td>
</tr>
<tr>
<td>HF</td>
<td>High frequency</td>
</tr>
<tr>
<td>HIV</td>
<td>Human immunodeficiency virus</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Description</td>
</tr>
<tr>
<td>--------------</td>
<td>------------------------------------</td>
</tr>
<tr>
<td>IC</td>
<td>Integrated circuit</td>
</tr>
<tr>
<td>IPT</td>
<td>Inductive power transfer</td>
</tr>
<tr>
<td>IR</td>
<td>Infra-red</td>
</tr>
<tr>
<td>LED</td>
<td>Light emitting diode</td>
</tr>
<tr>
<td>LF</td>
<td>Low frequency</td>
</tr>
<tr>
<td>LoC</td>
<td>Lab-on-a-chip</td>
</tr>
<tr>
<td>MEMS</td>
<td>Micro electro mechanical systems</td>
</tr>
<tr>
<td>MMF</td>
<td>Magnetomotive force</td>
</tr>
<tr>
<td>MOSFET</td>
<td>Metal oxide field effect transistor</td>
</tr>
<tr>
<td>NFC</td>
<td>Near-field communication</td>
</tr>
<tr>
<td>PAFC</td>
<td>Phosphoric acid fuel cell</td>
</tr>
<tr>
<td>PCB</td>
<td>Printed circuit board</td>
</tr>
<tr>
<td>PCR</td>
<td>Polymerase chain reaction</td>
</tr>
<tr>
<td>PEMFC</td>
<td>Proton exchange membrane fuel cells</td>
</tr>
<tr>
<td>PID</td>
<td>Proportional integral derivative</td>
</tr>
<tr>
<td>PMA</td>
<td>Power Matters Alliance</td>
</tr>
<tr>
<td>POC</td>
<td>Point-of-care</td>
</tr>
<tr>
<td>PWM</td>
<td>Pulse width modulation</td>
</tr>
<tr>
<td>RF</td>
<td>Radio frequency</td>
</tr>
<tr>
<td>RFID</td>
<td>Radio frequency identification</td>
</tr>
<tr>
<td>RMS</td>
<td>Root mean square</td>
</tr>
<tr>
<td>RNA</td>
<td>Ribonucleic acid</td>
</tr>
<tr>
<td>SAFC</td>
<td>Solid acid fuel cell</td>
</tr>
<tr>
<td>SHF</td>
<td>Super high frequency</td>
</tr>
<tr>
<td>SMA</td>
<td>Shape memory alloy</td>
</tr>
<tr>
<td>SMD</td>
<td>Surface mount device</td>
</tr>
</tbody>
</table>
UHF - Ultra high frequency
UPT - Ultrasonic power transfer
VLF - Very low frequency
WPC - Wireless power consortium
WPT - Wireless power transfer
ZCS - Zero current switching
ZVS - Zero voltage switching
μTAS - Micro total analysis systems

Symbols

μ - Magnetic permeability of the operating medium
μ₀ - Magnetic permeability of vacuum
μ_r - Relative permeability of material
A - Electrode surface area
A - Cross sectional area
AlNiCo - Aluminium Nickel Cobalt
B - Magnetic flux density
B - Magnetic flux density
B_{gap} - Magnetic flux density of airgap
B_r - Residual magnetism or remanence
B_{r_AlNiCo} - Remanence of AlNiCo
B_{r_NdFeB} - Remanence of NdFeB
C - Capacitance
C - Capacitance
C_p - parallel connected capacitor
Cs  - Secondary compensation capacitor

d  - Separation distance

D  - Diameter

D  - Spacing between conductors

E  - Energy density

E  - Energy

E  - Energy required

E_{density}  - Energy density

F  - Force

f_0  - Operating frequency

f_0  - Operating frequency/Resonant frequency

f_c  - Centre frequency of filter

f_{control}  - Control frequency

F_{gap}  - Force in airgap

f_P  - Power transfer frequency

h  - Length of coil

H  - Magnetic field strength

H_{op}  - Applied magnetic field

H_c  - Coercivity or coercive force

I  - Current

I_{act}  - Output current of energy buffer

I_{DC}  - DC input current

I_{DC}  - DC source input current

i_{DS1}  - Drain source current of switch 1

i_{DS2}  - Drain source current of switch 2

I_P  - Primary resonant current
<table>
<thead>
<tr>
<th>Symbol</th>
<th>Term</th>
</tr>
</thead>
<tbody>
<tr>
<td>$I_P$</td>
<td>Primary current</td>
</tr>
<tr>
<td>$I_R$</td>
<td>Maximum output current</td>
</tr>
<tr>
<td>$I_{sc}$</td>
<td>Short circuit current</td>
</tr>
<tr>
<td>$I_T$</td>
<td>Current through the resonant tank</td>
</tr>
<tr>
<td>$K$</td>
<td>Constant of the natural response</td>
</tr>
<tr>
<td>$k$</td>
<td>Coupling coefficient</td>
</tr>
<tr>
<td>$k_f$</td>
<td>Coupling factor</td>
</tr>
<tr>
<td>$l$</td>
<td>Total length of the copper wire</td>
</tr>
<tr>
<td>$L$</td>
<td>Length</td>
</tr>
<tr>
<td>$L$</td>
<td>Inductance</td>
</tr>
<tr>
<td>$L_{coil}$</td>
<td>Inductance of coil</td>
</tr>
<tr>
<td>Li-ion</td>
<td>Lithium Ion</td>
</tr>
<tr>
<td>$L_{in}$</td>
<td>Equivalent DC inductance</td>
</tr>
<tr>
<td>$L_p$</td>
<td>Primary inductance</td>
</tr>
<tr>
<td>$L_s$</td>
<td>Secondary pick up coil inductance</td>
</tr>
<tr>
<td>$L_s$</td>
<td>Secondary inductance</td>
</tr>
<tr>
<td>$m$</td>
<td>Maximum mass</td>
</tr>
<tr>
<td>$M$</td>
<td>Mutual inductance</td>
</tr>
<tr>
<td>$M_m$</td>
<td>Magnetization</td>
</tr>
<tr>
<td>$MMF_{AlNiCo}$</td>
<td>Magnetomotive force of AlNiCo</td>
</tr>
<tr>
<td>$MMF_{eq}$</td>
<td>Equivalent magnetomotive force</td>
</tr>
<tr>
<td>$MMF_{NdFeB}$</td>
<td>Magnetomotive force of NdFeB</td>
</tr>
<tr>
<td>$n$</td>
<td>Number of coil layers</td>
</tr>
<tr>
<td>$N$</td>
<td>Number of turns</td>
</tr>
<tr>
<td>NdFeB</td>
<td>Neodymium Iron Boron</td>
</tr>
<tr>
<td>Ni-Cd</td>
<td>Nickel Cadmium</td>
</tr>
<tr>
<td>Symbol</td>
<td>Description</td>
</tr>
<tr>
<td>--------</td>
<td>-------------</td>
</tr>
<tr>
<td>NiFe</td>
<td>Nickel Iron</td>
</tr>
<tr>
<td>Ni-MH</td>
<td>Nickel Metal Hydride</td>
</tr>
<tr>
<td>$N_p$</td>
<td>Number of turns in primary inductors</td>
</tr>
<tr>
<td>$N_s$</td>
<td>Number of turns in secondary inductors</td>
</tr>
<tr>
<td>$n_s$</td>
<td>Ratio between primary and secondary number of turns</td>
</tr>
<tr>
<td>$P_{act}$</td>
<td>Power delivered from energy buffer to actuator</td>
</tr>
<tr>
<td>$P_c$</td>
<td>Core losses</td>
</tr>
<tr>
<td>$P_{density}$</td>
<td>Power density</td>
</tr>
<tr>
<td>$P_m$</td>
<td>Maximum power transferred</td>
</tr>
<tr>
<td>$P_{max}$</td>
<td>Maximum power transferrable</td>
</tr>
<tr>
<td>$P_s$</td>
<td>Eddy current losses</td>
</tr>
<tr>
<td>$P_w$</td>
<td>Conduction losses</td>
</tr>
<tr>
<td>$Q$</td>
<td>Quality factor</td>
</tr>
<tr>
<td>$Q_p$</td>
<td>Primary quality factors</td>
</tr>
<tr>
<td>$Q_s$</td>
<td>Secondary quality factors</td>
</tr>
<tr>
<td>$r$</td>
<td>Radius of AlNiCo/NdFeB</td>
</tr>
<tr>
<td>$R$</td>
<td>Resistance</td>
</tr>
<tr>
<td>$R_{act}$</td>
<td>Resistance seen at actuator</td>
</tr>
<tr>
<td>$\Re_{AlNiCo}$</td>
<td>Reluctance of AlNiCo</td>
</tr>
<tr>
<td>$R_{coil}$</td>
<td>Resistance of coil</td>
</tr>
<tr>
<td>$\Re_{eq}$</td>
<td>Equivalent reluctance</td>
</tr>
<tr>
<td>$R_{ESR}$</td>
<td>Equivalent series resistance of supercapacitor</td>
</tr>
<tr>
<td>$\Re_{gap}$</td>
<td>Reluctance of airgap</td>
</tr>
<tr>
<td>$\Re_{iron}$</td>
<td>Reluctance of iron</td>
</tr>
<tr>
<td>$\Re_{NdFeB}$</td>
<td>Reluctance of NdFeB</td>
</tr>
<tr>
<td>$R_p$</td>
<td>Series connected variable resistor</td>
</tr>
</tbody>
</table>
$s$ - Displacement

SmCo - Samarium Cobalt

$T$ - Period

$t$ - Time

$t$ - Time

$t_0$ - Start up time

$T_s$ - Switching period

$V$ - Supply voltage

$V$ - Actuator volume

$V_{A(peak)}$ - Peak voltage across a switch at point A

$V_{AB}$ - Voltage across primary track inductor

$V_{act}$ - Voltage of actuator

$V_B$ - Resonant tank voltage at point B

$V_{B(peak)}$ - Peak voltage across a switch at point B

$V_C$ - Resonant voltage across capacitor

$V_{ctrms}$ - RMS voltage across capacitor

$V_{DC}$ - DC source voltage

$V_{Gi}$ - Gate voltage of switch 1

$V_{LP}$ - Voltages across primary inductor

$V_{oc}$ - Open circuit voltage

$V_P$ - Primary resonant voltage

$V_R$ - Maximum output voltage

$V_{RP}$ - Voltages across primary resistor

$V_{super}$ - Ideal source voltage of supercapacitor

$V_z$ - Forward Zener diode voltage

$w$ - Diameter of coil wire
<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$W$</td>
<td>Work delivered</td>
</tr>
<tr>
<td>$\varepsilon$</td>
<td>Permittivity of dielectric medium</td>
</tr>
<tr>
<td>$\eta_{\text{max}}$</td>
<td>Maximum efficiency</td>
</tr>
<tr>
<td>$\theta_i$</td>
<td>Initial phase angle of current</td>
</tr>
<tr>
<td>$\theta_v$</td>
<td>Initial phase angle of the voltage</td>
</tr>
<tr>
<td>$\rho_{\text{copper}}$</td>
<td>Resistivity of copper</td>
</tr>
<tr>
<td>$\tau$</td>
<td>T time constant</td>
</tr>
<tr>
<td>$\Phi$</td>
<td>Magnetic flux</td>
</tr>
<tr>
<td>$\Phi_{\text{AlNiCo}}$</td>
<td>Flux of AlNiCo</td>
</tr>
<tr>
<td>$\Phi_{\text{NdFeB}}$</td>
<td>Flux of NdFeB</td>
</tr>
<tr>
<td>$\Phi_{\text{total}}$</td>
<td>Net flux</td>
</tr>
<tr>
<td>$\chi_m$</td>
<td>Magnetic susceptibility of medium</td>
</tr>
<tr>
<td>$\omega$</td>
<td>Angular operating frequency / Zero voltage switching frequency</td>
</tr>
<tr>
<td>$\omega_0$</td>
<td>Angular resonant frequency</td>
</tr>
</tbody>
</table>
1.1. Microfluidic Systems

Microfluidic technology has gained much popularity in the recent years in biology, chemistry, medicine and more recently in various industries to overcome analytical challenges [1]. Microfluidic systems, transport and control infinitesimal quantities of fluids, where the fluid is pumped, mixed and manipulated through a micro-channel typically varying between 1-500 µm in width [2], for analysis at particle level [3].

In the micro-scale handling fluids is significantly different than in macro-scale since the fluid flow becomes laminar. Certain fluidic effects and surface forces such as surface tension [2], energy dissipation, and fluidic resistance, which are negligible in the macro-scale, become significant in the micro-scale [4]. Therefore, microfluidic systems are specifically designed for handling fluids in the micro domain.

These miniaturised microfluidic systems, which integrate chemical and biological processes on to a single platform, are commonly known as Lab-on-a-chip (LoC) systems or micro total analysis systems (μTAS). The key features of these are their small size, small volume, low energy consumption and utilisation of fluidic effects in the microdomain. LoCs allow fast, high throughput, parallel experiments using very limited reagent quantities [2]. LoCs are fabricated on silicon, polymers, glass or paper [3]. Some systems have internally integrated microvalves, micromixers and micropumps, to control and manipulate the flow of fluids, while in other systems these microfluidic components
are located external to the LoC.

LoCs are classified as channel based, droplet based and digital microfluidics [5]. Channel based microfluidics [6] have continuous streams of fluids processed in microchannels, as shown in Figure 1.1 [6-8].

![Figure 1.1: Channel based microfluidic systems [6-8]](image)

Droplet microfluidics have a sequential stream of droplets constrained within microchannels, shown in Figure 1.2 [6, 9, 10].

![Figure 1.2: Droplet based microfluidic systems [6, 9, 10]](image)

In digital microfluidics, shown in Figure 1.3 [11, 12], droplets of fluids are transported on Teflon surfaces with a pattern on electrodes beneath [1]. The voltages of these
electrodes are controlled systematically to drive the droplets across the surface, based on capacitive charge [1].

Figure 1.3: Microfluidic Systems [6, 11, 12]

1.1.1. Microfluidic Applications

Typical microfluidic tools vary among applications such as drug delivery, chemical synthesis, protein crystallization, cell cultures, point-of-care diagnostics, genetic sequencing, drug discovery etc. [2]. Microfluidic systems can handle bio-molecules such as cells, DNA, RNA, proteins or neurons and have found applications in to polymerase chain reaction (PCR) [13], DNA analysis and sequencing [14], protein separation [15], immunoassay and cellular analysis [16]. Today microfluidics is also popular in industrial and research applications in fluid dynamics, life sciences, chemistry, pharmaceuticals, biology and most engineering disciplines [2]. Example uses of microfluidics found in DNA amplification, immunoassays, cell-based assays, drug delivery and point of care diagnosis is shown in Figure 1.4 [17-20].

Creating multiple replicates of DNA sequences is known as DNA amplification [2]. Immunoassays are biochemical tests that measure the presence and concentration of analytes (typically proteins) in solutions using antibodies or immunoglobulin [2]. Assessing the effects of chemical stimuli on biological cells, especially in the
pharmaceutical industry for drug discovery, in basic biological research and chemotaxis studies is known as cell-based assaying [2]. Using microfluidic systems for these purposes have many benefits such as the reduction of reagent and sample consumption, processing time and fabrications costs. This in turn improves the portability of such devices. It also enables many parallel analyses and hence high throughput to improve data quality [2]. Microfluidic systems also help provide better control of test environments. The smaller dimensions of the cell and microchannels, laminar flow in channels, reduced dilution of analytes, and scaling down the electrical and magnetic fields are also some key advantages of microfluidic systems [2].

Figure 1.4: Real-life applications of microfluidic systems [17-20]

Conventional drug delivery methods can sometimes be inefficient due to long term treatment, complex dosages, combination of drugs and unstable active ingredients. Microfluidics in drug delivery can precisely control the amount of drug release, deliver drugs on demand in multiple doses, eliminate frequent injections and protect active ingredients [2].
Microfluidic technologies have improved point-of-care (POC) diagnostics by reducing reagent volumes, cost and analysis times, while enhancing the compact and portable nature of these devices. This facilitates suitable and prompt treatment, ensuring safe blood banking and improving clinical outcomes. Microfluidic systems can be used for POC diagnostics to analyse HIV infected blood samples, detect cancerous cells in blood components and separation of blood constituents [2].

Zebra fish embryos are often used in bio-analytical microfluidic systems, where the embryos are closely monitored under certain fluidic conditions. The automated biochemical laboratory [21], developed at the University of Auckland, is a classic example of the use of microfluidics for embryo monitoring, which will be discussed in detail in section 1.2.

Microfluidics is gaining popularity in more industrial applications such as the oil and gas industry where microfluidics is adopted for improved oil and gas analysis, property measurements and enhanced oil recovery [22].

1.1.2. Microfluidic Components

LoCs often need to work in conjunction with other microfluidic devices, tubing, connectors etc. to perform their tasks. Micropumps, microvalves, micromixers, separators and concentrators are utilised to move, mix, separate and process the micro fluid quantities.

Micropumps, shown in Figure 1. 5 (a) [23-26], create a pressure difference in the flow path to move sample fluids and reagents so that fluids can be directed in one direction or another. Mechanical pumps displace a diaphragm or flexible membrane to apply force to the fluid [27]. Non-mechanical pumps convert non-mechanical energy to kinetic energy to supply momentum to the fluid [28]. Microvalves are miniaturised valves used for fluid switching and control, as illustrated in Figure 1. 5 (b) [24, 25, 29]. In actively driven mechanical microvalves which are mainly built using micro electro mechanical
systems (MEMS), where a membrane is deflected when the actuating source is turned on [28].

Mixing in the microscale is far more difficult than in the macro-scale because flows are laminar and cannot use the turbulence of flow to mix as in the macroscale. Microfluidic mixers reduce the time for mixing. Passive mixers with no moving parts are cheaper while active mixers contain moving parts [30]. Several types of micromixers are illustrated in Figure 1. 5 (c) [25, 31].

![Figure 1. 5: Microfluidic flow control components](image)

1.2. Automated Bio-Chemical Laboratory

This research was inspired by some of the practical difficulties experienced by the Automated Bio-chemical Laboratory (ABL), which is a fully automated microfluidic platform developed to investigate the effects of toxicity on Zebra fish embryos [21]. It is designed around a LoC [32] which allows high throughput toxicity screening system for Zebra fish embryos, without human intervention [21]. The on-board LoCs were developed by the BioMEMS research group at the Department of Chemistry and the control of the platform was developed by the Department of Electrical and Computer Engineering, at the University of Auckland. The embryos are entrapped and observed subsequently using video technology [21]. The LoC consists of multiple fluid channels.
which is first loaded with embryos and chemicals and then used to observe and compare embryo development under exposure to different chemicals [21].

Overall, the ABL system is responsible for the following actions:

- Fluidics control
- Automated image acquisition
- Providing a graphical user interface
- Automated embryo loading
- Automated embryo sorting
- Embryo microenvironment control
- Perform different microscopy techniques
- Image processing
- Real-time data display

The ABL platform is made up of two distinct units: the embedded microfluidic unit, with its key components illustrated in Figure 1.6 [21], and the laboratory control unit.

![Figure 1.6: Automated Bio-chemical Laboratory (ABL)](image)
The embedded microfluidic unit consists of several peristaltic pumps, microvalves, a LoC platform, position control motor and an embedded microcontroller. The LoC can be moved in very fine steps, using a stepper motor, to place each section of the LoC under a high-resolution CMOS camera and acquire images of the embryos [21]. The laboratory control unit controls the overall functionality of the embedded unit. It acquires images of the embryos periodically which are loaded onto a computer for immediate and, post processing and analysis [21]. This unit can supervise multiple microfluidic units, allowing to create a networked sensing and control system [21]. Some photographs of the ABL platform are shown in Figure 1.7.

![Figure 1.7: Photographs of ABL platform](image)

Each experiment on the ABL platform consists of four stages: purging, loading, experiment and termination [21]. The experimental run begins with the purging state where the LoC is purged with clean water to remove any toxic residue from previous experiments. Both micropumps operate at full speed to circulate a large amount of water
through the platform. Valve 2 is opened to allow water into the LoC while the other valves remain closed.

The purging stage is followed by the loading stage, where an embryo specimen is loaded to fixed positions on the LoC every two seconds by opening valve 3. A constant flow rate and direction is maintained by operating pump 1 at a very low speed. Pump 2 is used to immobilise the specimens in small entrapments within the LoC channel.

Third is the experiment stage, where valve 1 and 4 are opened to flow the chemical through the channel and recycle it. Pump 1 and 2 operate at a constant low speed to maintain the flow rate of the chemical. The on-board camera photographs each section of the LoC channel to capture every single embryo, and the photographing process is repeated every 30 minutes. The screening process continues till the Zebra fish embryos are fully developed which typically takes up to three days. The human operator has control in specifying different termination criteria and experiment completion times over the laboratory control unit.

When the pre-organised termination time or termination criterion is reached the ABL platform will enter the termination stage. Pump 1 will start running at full speed in the reverse direction, with valve 2 opened for an inflow of water, to dispose the embryos. Pump 2 is disengaged to allow the specimens to move through freely.

1.2.1. Microfluidic Valves in ABL Platform

The ABL platform utilises four normally open solenoid pinch valves manufactured by Biochem™ microvalves [24] for the aforementioned operations. These valves have an actuator which is excited with a magnetisation current, to drive a solenoid to the closed position. When these valves are kept closed the current is drawn continuously leading to a significant amount of power being consumed.
Each valve consumes 2.9W of power when held in the closed position. The valves on the ABL platform require regular, infrequent actuation where valves 2 and 3 are kept closed for the entire experiment stage lasting up to 3 days. Valve 2 is only opened during purging and termination stages, which lasts for roughly 2 and 5 minutes respectively. Valve 3 is only opened during the loading stage for about 10 minutes during each full experiment run.

The actuation mechanism of the microactuator, is very important in terms of the power consumption of a microvalve. It is evident that improvements in the microactuator design for these valves could potentially reduce the power consumption of the ABL platform.

### 1.2.2. Requirement of a Novel Valve Actuator

With several valves staying closed for long hours, the power consumption is very high, leading to a very large energy requirement. A very large energy source is needed to meet this requirement which hinders the flexibility and portability of the ABL platform. A more energy efficient actuation mechanism is vital in making the system portable.

The continuous powering of the microvalves for prolonged periods causes significant heating in the ABL platform. This causes a significant increase in the temperature, known as Joule heating, which can potentially affect the fluidic properties and harm the Zebra fish embryos [33].

Wired power connectors for the microvalves in the fluidic environment is electrically unsafe as it can cause short circuits and chaotic. Accumulation of bacteria and other contaminants at these connection points is highly undesirable and would be good to avoid. With multiple pumps, valves and fluidic tubes on board the ABL platform, it poses a chaotic and unsafe environment. A wirelessly powered microvalve is highly desired in these type of environments, to improve the overall safety, cleanliness and allow a much higher degree of flexibility. Wireless power capability, on a microvalve, which will be
briefly introduced in section 1.3, also extends the valves versatility to the implantable devices arena.

Investigating the microfluidics market both in commercial and research level products proved that very little work had been conducted in developing energy efficient devices. Although there are many different valve actuation mechanisms in literature almost all require continuous power during operation, except for bistable valves [34-39]. Very limited work has been conducted in wirelessly powering and actuating [40-42] microfluidic devices. A thorough review of microvalves and their actuation mechanisms has been conducted and presented in Chapter 2, it shows that a single microvalve actuator addressing the aforementioned concerns has not been developed to date. Therefore, an energy efficient, wireless microfluidic valve is highly sought after for the ABL platform discussed above and other similar microfluidic platforms.

In order to create a more energy efficient and flexible state-of-the-art solution, an energy efficient, wireless valve actuator, referred to as the microactuator, needs to be developed. The ABL platform is the primary target of the microactuator proposed in this thesis, although it can easily be adapted for other microfluidic platforms.

1.3. Wireless Power Supplies

As wireless power is a primary focus of the novel microactuator, it is important to investigate wireless power supply options currently available. Wireless power transmission (WPT) collectively refers to the wire-free transfer of electrical power using electromagnetic waves. The concept of WPT was brought forward by Nikola Tesla at the start of the last century [43, 44]. The basic concept is based on a power source transmitting field energy across a space to a single or multiple power receivers, where the energy is converted back to an electrical current to power target devices. It is becoming increasingly popular in consumer electronics and electric vehicle (EV) charging. Wireless power ICs by Powercast, IDT, Texas Instruments, Toshiba, NXP etc. are currently
available commercially. One of the most popular and oldest wirelessly powered devices is the electric toothbrush. The wireless charging mat for mobile devices, is now available through Samsung, Microsoft, Power-by-Proxi etc. as seen in Figure 1. 8. The waterproof wireless power connectors developed by Power-by-Proxi are also illustrated in Figure 1. 8.

![Figure 1.8: Commercially available wireless power devices [45, 46]](image)

With the introduction of EVs, both static and dynamic wireless charging of EVs has become a very popular research area. EV manufacturers such as Tesla, BMW, Mercedes, Toyota, Audi, Nissan, Honda and General Motors work together with wireless power specialists such as Qualcomm, Evatran, Toshiba to introduce wireless charging into their upcoming products. This is a major step towards popularising greener EVs in place of fossil fuel driven vehicles. Figure 1. 9 shows several EVs being charged wirelessly both in static mode and dynamically via embedded in-road charging pads [47-49].

![Figure 1.9: Wireless charging of EVs when stationary and on the road [47-49]](image)
Another major application area of WPT is powering implanted prosthetic devices in the human body, to avoid wires passing through the skin [50]. Currently most research is focused on wirelessly powering implantable devices, ranging from cochlear implants (hearing aids) to heart assist devices (pace makers). Much research is underway for charging cardiac pacemakers, insulin pumps and various other bio-medical implants. Wirelessly powering microfluidic components has not yet been investigated but has much potential as well as challenges due to the compact and portable nature of these systems.

Wireless power can be facilitated through inductive power transfer (IPT), capacitive power transfer (CPT), ultrasonic power transfer (UPT), laser and microwave. IPT is the most mature technology and is widely used in many consumer applications. Wireless power allows safe, spark free operation, no mechanical wear and tear, and a much higher degree of mechanical freedom. The main trade-off in WPT is the lower efficiency in comparison with 100% efficient wired connector. Further details on wireless power are discussed in Chapter 5.

To enhance the wireless power capability of the system, energy buffering can be utilised to temporarily store energy onboard. This allows for energy to be stored when the wireless power link is available and use the same energy to meet peak power requirements of the application.

1.4. Objectives and Scope of Research

The objective of this research is to develop a microvalve actuator which can address the need for a more flexible and energy efficient solution. After reviewing existing literature, presented in Chapter 2, it has been found that there is much room for improvement in energy efficient design of microvalve actuators. Actuation mechanisms of existing valves were investigated in detail to design a novel actuation mechanism which is energy
efficiency, and better physical performance and flexibility, and therefore better suited to drive the valves on the ABL platform.

This research focuses on delivering a bi-stable actuator that features the following:

- Energy efficient bi-stable actuation technique
- Wirelessly powered microactuator
- Energy buffering for offline operation of the valve when power is unavailable
- Wirelessly controlled actuation

The scope of this research includes the development of a novel, state-of-the-art actuation mechanism which significantly reduces the energy consumption of the microvalve actuator. The proposed actuation mechanism is completely novel to the microfluidics arena, and very versatile with potential actuator applications in other areas too.

Existing wireless power supply technologies are investigated to design a suitable wireless power solution for the microactuator. The primary side of the wireless power supply will work in conjunction with the microfluidic controller and with the secondary circuit enclosed within the microactuator. The wireless power solution will have the ability to operate offline with the aid of on-board energy storage which also meets peak power requirements of the microactuator. Research on microfluidic valves has shown that wireless power is a highly anticipated feature which is not found on any microfluidic valve actuator to date.

With the removal of wired power connections to the microactuator, a wireless control signal communication mechanism will also be developed as part of this research. The proposed technique will use minimal additional resources to keep the power consumption and real estate of the actuator to a minimum. This technique too is very unique and efficiently utilises existing wireless power supply resources for wireless actuation control.
1.5. Thesis Organisation

The remaining chapters are organised as follows. Chapter 2 reviews existing literature in microvalves, their key performance parameters and electromagnetically driven valve types. The solenoid valves used on the target microfluidic platform are introduced with their drawbacks highlighted. A review of state-of-the-art microvalves which offer bi-stable actuation and wireless power is presented. These valves are the best performers in terms of energy efficiency and physical performance and their key performance parameters have been identified and summarised.

Chapter 3 presents the proposed microvalve actuator and its key features which include, ultra-energy efficient bi-stable actuation mechanism, wireless power supply, onboard energy buffering and wireless control signal communication.

Chapter 4 discusses details of the actuation mechanism, outlining the concept, material selection process and basic mechanical design. Magnetic simulations of the actuator have been conducted using FEMM modelling software. A thorough analysis of the magnetic and electrical circuits has also been conducted. Design and implementation details of the magnetic excitation and position feedback circuits are discussed. Experimental results highlighting the power and energy performance, actuation speed and physical capabilities of the novel actuation technique are presented.

Chapter 5 details the power supply design which includes wireless power and energy buffering. Existing wireless power supply techniques are summarised to outline their different attributes and justify the selection of IPT technology. The basic operating principles of IPT, the setup of the IPT supply in the microactuator application, design and implementation details of the primary and secondary wireless power circuits are discussed. Energy storage options currently available for energy buffering are summarised to highlight supercapacitor technology as the most feasible choice. The design details of the energy buffer are presented. Experimental results highlighting the power supply performance, and energy storage capability are presented.
Chapter 6 details the wireless control strategy used in the actuator. Short range wireless communication techniques are summarised. The need for a simpler communication technique using existing resources to avoid design complexity, increased power consumption, real estate and cost is shown. Design and implementation details of the developed wireless communication technique are given with experimental results highlighting its success in this context.

Chapter 7 discusses the overall performance of the full actuator and summarises the overall performance of the novel microactuator. The energy performance of this actuator is compared with the solenoid valves on the target microfluidic platform, and with other active microfluidic valve actuators and the highest performing state-of-the-art microvalves with bi-stable actuation.

Chapter 8 draws conclusions and makes recommendations for future work to improve the proposed wireless microactuator design.
Chapter 2

Literature Review

2.1. Microvalves

A microvalve is a mechanical device that contains pressure to block or modify the flow of a fluid that passes through it [51]. They are used to control routing, timing and separation of fluids in microfluidic systems [52]. Microvalves are manufactured in a variety of sizes ranging from 50 µm up to 1mm for orifice diameters [53] and use a vast range of actuation mechanisms. Microvalves are broadly classified as active or passive devices, depending on the requirement of external power to actuate the valve [54]. Passive valves are easy to fabricate, require no external energy to operate and are easily integrated into microfluidic networks with the downside being their limited functionality and diversity [54]. Active valves, also known as ‘intelligent valves,’ require external energy to operate additional actuators but allow more advanced control of the fluid flow [30]. Active microvalves rely on actuation energy, such as electric, magnetic, electrostatic, piezoelectric, bimetallic or pneumatic energy [2].

Since their invention in 1979 [55] microvalves have evolved into many different designs and actuation schemes. Generally, a microvalve consists of three main parts, the valve seat, the membrane/diaphragm and the actuator [56]. The valve seat interfaces with the rest of the microfluidic system and makes up the base of the valve [56]. The actuator provides the necessary force to the membrane which deflects it to close or open the valve [56]. The valve seat is driven by an actuator to change the position of the valve [51]. The valve body contains the fluid and its pressure while the actuator controls the position of
the valve seat [51]. The means by which this flexible diaphragm is compressed to close the valve is known as the actuation mechanism.

While there are many ways to classify microvalves, one way is based on their initial working state as normally open, normally closed, and bistable. Bistable microvalves allow the valve to be latched in either the open or closed positions and do not have a defined off-state [51].

The performance and size of active microvalves are dependent on the actuator, which can be broadly classified as miniaturized external and micromachined [54].

Microvalves can be classified according to the actuation method, as shown Figure 2.1 [57], but are also classified based on their energy requirement, normal operation state, type of fluids handled etc. [52, 58].

![Classification of microvalves](image)

*Figure 2.1: Classification of microvalves*

Externally actuated active microvalves such as electromagnetic, piezoelectric, pneumatic and Shape Memory Alloy (SMA) provide extra control over the fluid flow. Figure 2.2 illustrates the operating mechanism of the common externally actuated valve types [57].
A solenoid plunger, sometimes combined with permanent magnets to increase magnetic forces, compresses the membrane in electromagnetic actuation [57]. An applied electric field enables piezoelectric actuations to deflect the microvalve membrane [57] as seen in Figure 2.2 (b). Pneumatic microvalves, also commonly referred to as Quake valves, use an additional fluid channel perpendicular to the flow channel to block or release the fluid flow [59]. Pneumatic valves are popular in high-temperature applications because the operating temperature range depends solely on their material [51]. SMA, an alloy which remembers its original state, has a cold-forged shape. When heated they change shape to actuate the microvalve by returning to their pre-deformed shape [60].

The major specifications that define the performance for any type of microvalve are leakage, valve capacity, power consumption, closing force (pressure range), temperature range, response time, reliability, biocompatibility, and chemical compatibility [51]. Ideally the active valve should have zero leakage in the closed position [51]. The valve capacity refers to the maximum flow rate the valve can handle, while the closing force depends on the pressure generated by the actuator. The material and actuation mechanism strongly define the valves operating temperature range. The response time of the actuator refers to the valve response time [51].

The power consumption of the active microvalves is the main focus in this research. By definition, the valve’s power consumption is the total input power of the valve in its active, power-consuming state [51]. Based on the actuating principle, the power consumption varies by several orders of magnitude [51].

Figure 2.2: Externally actuated active microvalve mechanisms [57]
2.1.1. Performance Parameters of Microvalve Actuators

In active microvalves the external or micromachined actuator plays a very important role. The actuator is responsible for the moving, holding and other dynamic functions of the microvalve [51]. The actuator should provide sufficient force, displacement and controllability to move the valve seat to the desired position, which is known as the moving function [51]. The actuator together with the valve spring will need to overcome the pressure from the fluid inlet to facilitate the holding function [51]. The dynamic function determines the response time of the valve [51].

The performance of the actuator is characterized by the work delivered $W$, which can be quantified by (2.1), where $F$ is the actuator force and $s$ is the maximum displacement [51].

$$W = F \cdot s \quad (2.1)$$

Since the size of the actuator is crucial for miniaturization and the energy density $E$ is usually used to rate an actuator’s performance. The energy density $E$ can be defined by (2.2), where $V$ is the actuator volume [51]. Figure 2.3 shows the range of energy densities found in several types of microvalve actuators [51].

$$E = \frac{W}{V} = \frac{F \cdot s}{V} \quad (2.2)$$

![Figure 2.3: Energy density range of microvalve actuators [51]](image)

Certain material parameters of each actuator type can be used to quantify the energy density. Particularly for electromagnetic actuators it is given by (2.3), where $B$ is the magnetic flux density, $\mu = \mu_r \mu_0$ is the permeability of the operating medium, $\mu_0 = 4\pi \times 10^{-7} = 1.26 \times 10^{-6}$ H/m is the permeability of vacuum, and $\mu_r$ is the relative permeability of the material [51].
Most microvalve actuators are dependent on an electrical input, but some types of bistable valves utilise chemical and potential energy [51]. In general, active microvalves, apart from latching bistable valves, require a continuous supply of power and the magnitude of this differs vastly based on the type and size of the actuator. The power consumption of the microfluidic components is directly proportional to the heat generated in a microfluidic platform, the size and type of power source required, the dimensions and portability of the microfluidic system, fluid reservoir volume [61] etc. Piezoelectric valves require the highest operating voltage range 10-500 V, while SMA valves need less than 1 V. Electromagnetic valves require voltage in the range of 3-30 V [34, 57, 62-68]. However, the power requirement is highest for electromagnetic valves which varies between 0.4-30 W. Lowest power consumption levels are observed in SMA valves which is around 90-400 mW. Although piezoelectric valves require very high operating voltages their power consumption is much lower around 1-500 mW [34, 57, 62-68]. Depending on the type of microvalve used the power requirement may from µW to W range, making it difficult to design a generalised solution for wirelessly powering all types of microvalves. Therefore, electromagnetic valves, being the highest power consumer, are further investigated to be redesigned with higher energy efficiency, equipped with wireless power.

2.1.2. Electromagnetic Microvalve Actuators

Electromagnetically actuated valves were chosen in the ABL platform for fluid flow control, as they offer a large deflection [51] and hence provide better control. The vertical force $F$ offered by the magnetic flux $B$ acting in the direction $z$ acting on a magnet with magnetisation $M_m$ and volume $V$, is given by (2.4) [51].

$$F = M_m \int \frac{dB}{dz} dV \quad (2.4)$$
The magnetic flux \( B \) and magnetic field strength \( H \) are related by (2.5), where \( \mu, \mu_0, \mu_r \) and \( \chi_m \) are the permeability, permeability of free space (\( 4\pi \times 10^{-7} \text{ H/m} \)), relative permeability and magnetic susceptibility of the medium, respectively [69].

\[
B = \mu H = \mu_r \mu_0 H = \mu_0 (1 + \chi_m)H
\]

(2.5)

Solenoid actuators are often found in electromagnetic microvalves, which have a magnetic core and coil to generate a magnetic field, which may be integrated on silicon for compactness [51]. The main drawback of these valves is the low energy efficiency due to heat loss in the coil. The magnetic field strength \( H \) in the coil with \( N \) turns, \( L \) long driven by a current \( I \) is given by (2.6) [69].

\[
H = \frac{NI}{L}
\]

(2.6)

Different electromagnetic actuator configurations are illustrated in Figure 2.4 [51].

---

**Figure 2.4: Electromagnetic actuator configurations [51]**
Figure 2.2 (a) shows a typical solenoid microvalve with an external actuator where the valve seat and nozzle are micromachined in silicon [51]. In the valve seen in Figure 2.2 (b) the valve seat is in the form of a plate suspended on four folded beams. The gap between the seat and nozzle is changed proportionally to operate the valve [51]. In this case the valve seat is made of NiFe, which serves as the magnetic core. The current carrying coil is external to the fluid tube, allowing electrical isolation [51]. Figure 2.2 (c) also illustrates an externally actuated microvalve which can either keep the valve open or closed [51]. The tightness and the resistance against particles is improved by the rubber membrane [35]. Figure 2.2 (d) is equipped with an integrated micro-coil made of electroplated gold, placed on the valve seat and an external magnetic core [51]. The valve seat is suspended on four folded beams, and this valve can achieve a closing force of 800 μN with a current of 25 mA and a field gradient of 310 T/m [70]. A flapper valve seat rotating under magnetic force is shown in Figure 2.2 (e). The coil integrated in the valve body is made of NiFe while the magnetic core is made up of magnetic foil and the complete package is magnetised [36]. The micro-coil in the normally closed needle valve shown in Figure 2.2 (f) is micromachined and the valve is driven by a PWM signal which generates a proportional flow rate at a constant inlet pressure [51].

2.2. Solenoid Valves in ABL Platform

The microvalves utilised in the ABL platform are normally open solenoid pinch valves from BioChem™ microfluidics, for which the simplified internal structure is shown in Figure 2.5. A current is driven through the solenoid coil, which generates a magnetic field to retract the valve plunger. This opens the tubing of the normally closed valve while closing the normally open valves [24]. De-energizing the solenoid removes the magnetic attraction, allowing the spring to push the plunger back to its original position [24].
2.2.1. Solenoid Valve Operation in Experimental Run

The Bio-Chem™ normally open pinch valve used in the ABL system requires 2.9 W of continuous power during operation to stay in the closed position [24]. The power consumption of each valve during a single experimental run is outlined in Table 2.1.

Table 2.1: Power and energy consumption of microvalves in ABL platform

<table>
<thead>
<tr>
<th>Valve</th>
<th>Purging Stage</th>
<th>Loading Stage</th>
<th>Experimental Stage</th>
<th>Termination Stage</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Valve 1</strong></td>
<td>2 minutes</td>
<td>10 minutes</td>
<td>3x24x60 minutes</td>
<td>5 minutes</td>
</tr>
<tr>
<td>Closed</td>
<td>Closed</td>
<td>Open</td>
<td>closed</td>
<td></td>
</tr>
<tr>
<td>2.9W -&gt; 0.348 kJ</td>
<td>2.9W -&gt; 1.74 kJ</td>
<td>0W -&gt; 0J</td>
<td>2.9W -&gt; 0.87 kJ</td>
<td></td>
</tr>
<tr>
<td><strong>Valve 2</strong></td>
<td>2 minutes</td>
<td>10 minutes</td>
<td>3x24x60 minutes</td>
<td>5 minutes</td>
</tr>
<tr>
<td>Open</td>
<td>Closed</td>
<td>Closed</td>
<td>Open</td>
<td></td>
</tr>
<tr>
<td>0W -&gt; 0J</td>
<td>2.9W -&gt; 1.74 kJ</td>
<td>2.9W -&gt; 751.68kJ</td>
<td>0W -&gt; 0J</td>
<td></td>
</tr>
<tr>
<td><strong>Valve 3</strong></td>
<td>2 minutes</td>
<td>10 minutes</td>
<td>3x24x60 minutes</td>
<td>5 minutes</td>
</tr>
<tr>
<td>Closed</td>
<td>Open</td>
<td>Closed</td>
<td>Closed</td>
<td></td>
</tr>
<tr>
<td>2.9W -&gt; 0.348 kJ</td>
<td>0W -&gt; 0J</td>
<td>2.9W -&gt; 751.68kJ</td>
<td>2.9W -&gt; 0.87 kJ</td>
<td></td>
</tr>
<tr>
<td><strong>Valve 4</strong></td>
<td>2 minutes</td>
<td>10 minutes</td>
<td>3x24x60 minutes</td>
<td>5 minutes</td>
</tr>
<tr>
<td>Closed</td>
<td>Closed</td>
<td>Open</td>
<td>Closed</td>
<td></td>
</tr>
<tr>
<td>2.9W -&gt; 0.348 kJ</td>
<td>2.9W -&gt; 1.74 kJ</td>
<td>0W -&gt; 0J</td>
<td>2.9W -&gt; 0.87 kJ</td>
<td></td>
</tr>
<tr>
<td><strong>Total Energy</strong></td>
<td>1.044 kJ</td>
<td>5.22 kJ</td>
<td>1.50 MJ</td>
<td>7.83 kJ</td>
</tr>
</tbody>
</table>
The total energy consumed during a single experimental run can be as high as 1.514 MJ, and therefore would require a bulky power source and hence pose significant limitations on the systems portability. It is evident that the most energy is consumed to keep valve 2 and 3 closed during the experiment stage which extends over several days. A more energy efficient option to achieve this task is vital, for many reasons which will be discussed in the next section.

### 2.2.2. Drawbacks of Solenoid Valves

The solenoid valves currently in use have several drawbacks, which pose several limitations on the ABL platform, such as:

- Significantly higher power consumption
- Significant heating in the platform
- Chaotic and non-flexible environment

The main drawback observed with the solenoid valves is the significantly large power consumption. The continuous power drain requiring mega joule level energy makes it impossible to use a portable power source such as a battery. Although the ABL platform analyses fluids in the micro scale it requires energy in the mega scale. By reducing the energy requirement of the microvalves and hence the required battery capacity, portability of such microfluidic platforms can be significantly improved.

With mega joules of energy dissipated in these valves, a significant temperature increase is experienced on the platform. While the heat energy increases the external system temperature, part of this thermal energy will be conducted to the fluids under experimentation and may significantly elevate the system enthalpy [71]. The buffer temperature of the system may increase and give rise to a temperature gradient across the buffer solution [72]. The increase in temperature can also change fluid properties such as viscosity, dielectric constant, electric conductivity, mass diffusivity, electromobilities and pH value [72].
The solenoid valves, like all other active microvalves are powered via two wire connectors. The wired power connectors for four individually controlled valves bring forward issues such as [73]:

- Chaotic nature due to wires in addition to the fluidic tubes
- Limitations on mechanical flexibility and portability of the system
- Potential short circuit of electrical connectors due to the liquid environment
- Chemical compositions create electrical erosion causing potential interference with continuous power supply
- Mechanical wear and tear, electrical erosion of the wired connectors
- Harbouring of bacteria and other contaminants at wired connection points

Research and experience in this context goes on to show that there is a dire need and much demand for a novel microvalve actuator which can enhance the portability of a microfluidic system by improving its energy efficiency.

### 2.3. State-of-the-Art Microvalves

Among the many different microvalves that exist, there is no valve that combines the features proposed in this research into a single package. Bistable valves, also known as latching valves are the most energy efficient type, given that these valves only consume power during actuation and holding the current state of the valve is done by other mechanisms. Bistable valves are commonly actuated using electromagnetic, SMA, electrostatic or thermopneumatic techniques [74]. Some state-of-the-art valves that deliver energy efficient features will be summarised in this section, with their feasibility in the ABL platform assessed.

In general latching solenoid valves have a permanent magnet placed at the top of the valve, with a small magnetic field which is insufficient to activate the plunger. A solenoid is energised with full power to create a strong field to attract the plunger towards the permanent magnet. Once the plunger is attracted to the permanent magnet, the force
exerted by the power is discontinued and the solenoid unenergized, yet the permanent magnet’s force is strong enough to hold the plunger in position. Similarly, the solenoid is energised with a reverse polarity current to remove the plunger from the permanent magnet, where the magnetic field generated by the solenoid is much stronger than that of the permanent magnet to remove the plunger from the permanent magnet’s grasp.

These types of valves are especially beneficial in applications where valves remain open for prolonged periods or if the valve is battery powered. Most common latching type solenoid valves have a current pulsed for approximately 20-50 ms and precise control of this pulse is vital [37]. If the pulse is too short the valve will not actuate and if too long the valve will open and then close again or vice versa, known as re-latching [37]. A very stable supply voltage which does not vary more than 10-15% is important for stable operation [37]. These valves do not operate well under vibrating conditions [37]. The power requirement of these valve coils ranges from about 0.65 W to 9 W [37] causing each actuation to use up about 13 mJ to 450 mJ. One major limitation of latching valves is that they are not suitable as safety valves. If power to a normally open or closed valve is interrupted the valve will resort to its default normally open/closed position. This will not happen with non-latching valves, but since the current position is unknown a reboot will be required [37].

Sutanto, J. et al. 2007, discuss a bistable electromagnetic valve which combines a permanent magnet to latch the valve in open/close position while requiring power only during actuation [38]. While these valves require an actuation voltage under 5 V and current of 570 mA approximately, the deflection offered is in the μm range varying from 20-35 μm. Experimental results have also shown that the valves require pulses of approximately 500 μs, leading to an approximate energy consumption of 1.425 mJ per actuation. However, as these valves offer very small deflections, and therefore is not suitable in the ABL application.

Bohm, S. et al. 2000, [35] present a bi-stable magnetically actuated valve using a solenoid and permanent magnet combination as in [38]. This microvalve delivers attraction and holding forces of 2.15 N and 2.7 N, respectively. A current pulse of 500 mA/100 ms is
required to perform each actuation. As the wattage or supply voltage information is unavailable, the energy consumption per actuation cannot be deemed. This valve offers a deflection of 200 µm, which is significantly larger than that of [38].

Capanu, M. et al. 2000, also describe a similar bistable valve with a cantilever mechanism in [36]. The valve was opened with a 5.3 ms minimum pulse of 1.38 V/0.34 A consuming 2.48 mJ [36]. The practical operating range was slightly higher at 4.26 mJ per actuation.

Chang, P. J. et al. 2012, discuss a micro-ball valve which consumes as low as 50mW during its magnetic actuation process in [34]. Paramagnetic spheres are dispersed through the array of microchannels, where the valves magnetically capture a sphere to close the fluidic channel when actuated. Current pulses of opposite polarity, ranging between 3-8mA are used to open and close these valves. The microactuator delivers a force of 10-15 nN, which is sufficient in micromachined actuators [34]. In case of the ABL platform this force is insufficient, and the design mechanism is unsuitable for the specific application.

Shape memory alloy (SMA) based bistable valves also offer low power actuation where the valve is latched in position and power is not consumed to hold the state. SMAs, as previously mentioned, are alloys which have a memorised state they can return to when heat is applied.

Megnin, C. et al. 2012, describe a technique which uses two SMA microbridges which deform when selectively heated to toggle the valve position [74]. These are used in combination with permanent magnets which hold the position while the SMAs return to their original state when cooled. Deflections of up to 50 µm and a maximum switching force achievable is approximately 290 mN have been achieved [74]. Pulses of 500 mW are applied to switch the valve for 200 ms, which leads to a power consumption of 100 mJ per actuation [74].

Yang, B. et al. 2007, describe a novel technique using the phase change of paraffin wax to latch a microvalve in position, however the power consumption information is not given [39]. Heat is applied to melt paraffin, with additional pneumatic pressure to switch the valve from open to closed. Energy is used only to liquefy the paraffin which solidifies in
position when the heat is removed. Table 2.2 summarises the key parameters of the aforementioned actuators [34-39].

Table 2.2: Performance of low power latching valves

<table>
<thead>
<tr>
<th>Microvalve</th>
<th>Current (A)/ Voltage (V)</th>
<th>Power</th>
<th>Pulse</th>
<th>Actuation Energy</th>
<th>Force</th>
<th>Deflection</th>
</tr>
</thead>
<tbody>
<tr>
<td>Solenoid Solutions Inc.</td>
<td>- / 0.65-9W</td>
<td>20-50ms</td>
<td>13-450mJ</td>
<td>-</td>
<td></td>
<td>-</td>
</tr>
<tr>
<td>Sutanto, J. <em>et al.</em> (2007)</td>
<td>570mA &lt; 5V</td>
<td>2.85W</td>
<td>500µs</td>
<td>1.425mJ</td>
<td>-</td>
<td>20-35µm</td>
</tr>
<tr>
<td>Bohm, S. <em>et al.</em> (2000)</td>
<td>500mA</td>
<td>-</td>
<td>100ms</td>
<td>-</td>
<td>2.15-2.7N</td>
<td>200µm</td>
</tr>
<tr>
<td>Capanu, M. <em>et al.</em> (2000)</td>
<td>340mA 1.38V</td>
<td>-</td>
<td>5.3ms</td>
<td>2.48-4.26mJ</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Chang, P. J. <em>et al.</em> (2012)</td>
<td>3-8mA</td>
<td>50mW</td>
<td></td>
<td>10-15nN</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Megnin, C. <em>et al.</em> (2012)</td>
<td>-</td>
<td>500mW</td>
<td>200ms</td>
<td>290mN</td>
<td>50µm</td>
<td></td>
</tr>
</tbody>
</table>

Similarly, very limited work has been conducted in wirelessly powering and actuating microvalves. Baek, S-K., *et al.* 2013, describes a technique to wirelessly actuate a disposable paraffin wax based microvalve using inductive heating [40]. Tikka, A. C., *et al.* 2011, propose a near field inductive power transfer based technique to power a human implantable surface acoustic wave correlation based passive microvalve [41]. Mohamed Ali, M. S., *et al.* 2013, have developed a radio controlled microactuator using SMA spiral coil inductors [42]. In this system, the SMA coil forms the inductor of the LC tank that couples with the transmitter and produces Joule heating to deform itself to its flat remembered state [42].

Literature indicates that although specific aspects of the proposed microvalve actuator have been investigated on a research base, there is much potential for a microvalve of the proposed nature to be a highly successful commercial product. The proposed microactuator will be a ground breaker in a new class of bi-stable wireless microactuators.
Chapter 3

Proposed Bi-Stable Microactuator

This thesis proposes, develops and analyses a bi-stable microvalve actuator which uses minimal energy and is powered and controlled wirelessly. The novel design addresses the drawbacks highlighted in Chapter 1 and is different to those found in literature discussed in Chapter 2. The proposed actuator will deliver the following features outlined in Figure 3.1.

![Figure 3.1: Features of proposed microvalve actuator](image)

The proposed actuator has an ultra-energy efficient actuation technique and does not need nearly as much energy as any of the microvalves currently available in the market today. In the proposed valve actuator power is only consumed during actuation and not required to maintain the state of the valve, which clearly differentiates it from the conventional solenoid valve. This actuation mechanism differs from existing bistable
valves, as it does not use any of the conventional mechanical or permanent magnet based latching techniques previously discussed in section 2.3.

An additional feature that enhances the flexibility, freedom of mechanical movement and portability of the microfluidic platform is the addition of wireless power to the microactuator. A reliable, robust, energy efficient wireless power technique will be devised to power the entire microactuator circuit. The proposed microactuator is self-sustainable with an onboard energy buffering system large enough to source several actuations before recharging is required. The peak power requirement of the proposed actuation mechanism makes an onboard energy buffer a vital requirement.

As wireless power becomes an attribute of the microactuator, a wireless control technique is needed to deliver the control command to toggle the state of each microvalve. The actuator becomes even more unique with its utilization of the existing wireless power link for wireless control command delivery, without the addition of extra wireless communication modules.

The proposed microvalve actuator is a highly versatile and scalable solution which is easily adapted to support many different micro-scale actuation applications.

The primary side of the wireless power supply is attached to the ABL platform and is controlled mainly by the FPGA with inputs from the ABL controller for actuation control. The remaining components including the secondary wireless power pickup, energy storage, position feedback and actuation circuits along with the wireless control circuit will be packaged into the microvalve. An overview of the proposed microactuator is presented in Figure 3.2.
This can be further elaborated as shown in Figure 3. 3, where the ABL platform houses the IPT primary track and converter and supplies it with DC power. The controller of the ABL generates the control command for the microvalve, which is delivered over the IPT converter. The remaining modules are all housed within the microvalve, including the secondary pick-up coil.
3.1. Electropermanent Magnet Based Actuation

Electropermanent (EP) magnets, a combination of hard and semi-hard magnets, are used to actuate the valve. A current carrying coil is excited with a short pulse to alter the magnetic field of only the semi-hard magnet. When both hard and semi-hard magnets are magnetised in the same direction the total magnetic flux attracts the plunger and releases the valve for continuous fluid flow. When the magnetic field of the hard and semi-hard magnet oppose each other, the total flux is contained within the magnetic structure to release the plunger and compress the fluid flow. The direction of the current pulse determines the resulting magnetic field pattern. This concept is illustrated in Figure 3.4.

![Figure 3.4: Basic concept of electropermanent magnet actuation](image)

3.2. Wireless Power and Energy Buffer

The primary wireless power supply is controlled in conjunction with the microfluidic platform to generate AC power. An AC current running through the primary coil generates a varying magnetic field which is coupled with the secondary pick up circuit which extracts power. The wireless power mechanism chosen for this application is inductive power transfer (IPT). A high frequency switching converter and primary track, external to the microvalve, are attached to the ABL platform. The switching converter generates a rapidly alternating current in the primary track at a fixed frequency. The secondary pickup coil placed in close proximity, is tuned to the same frequency to
resonate with the primary track and transfer power. Power transfer between the primary track and secondary coil is similar to a transformer, with the key difference being the loose coupling in the IPT system. The pickup and other circuitry are housed within the actuator where the main challenge is fitting into limited real estate. The basic concept is outlined in Figure 3.5.

![Figure 3.5: Basic wireless power supply layout](image)

The secondary circuit is attached to the supercapacitor energy buffer which will charge itself whenever power is available. This allows the valve to store energy when power is available and operate offline if power becomes unavailable. This eliminates the need to always depend on the IPT link during operation. During actuation, a high current is required for a brief time which cannot be sourced directly by the IPT power supply and the energy buffer allows this peak power requirement to be easily met.

### 3.3. Wireless Control of Actuator

The same IPT link is employed to deliver the control signal to toggle the state of the actuator. The secondary wireless power circuit also serves as a receiver antenna for the wireless control signal to actuate the valve. This signal is differentiated by the control signal detection circuit to issue the actuation command to the actuator control circuit. This command controls the power output from the energy buffer to the actuation and position feedback circuits, sourcing energy for the actuation. The control signal triggers
a position feedback request which is returned to the actuation circuit to then deliver the required control pulse to toggle the state of the microvalve.

The control of the actuator is achieved using discrete semiconductors such as gates, timers, comparators and passive filters to avoid the use of microprocessors leading to continuous power consumption. The flow of the control logic is given in Figure 3.6.

![Diagram](image)

Figure 3.6: Logical flow of the microactuator control
Chapter 4

Bi-Stable Actuation Based on Electro-Permanent Magnets

The ultra-energy efficient, magnetic actuation mechanism is the most significant feature of the proposed microactuator. This novel technique uses a combination of magnetic materials to actuate the valve to facilitate bistable actuation.

4.1. Electro-Permanent Magnet

The actuation is performed using electro-permanent (EP) magnets, which combine features of electromagnets and permanent magnets. Figure 4.1 illustrates the basic actuation concept of the EP magnet actuator.
A hard-magnetic material is placed alongside a semi-hard magnetic material to form the EP magnet and the combined field generated by these magnets control the actuation. An external magnetic field is applied using a current carrying coil which changes the magnetisation direction of the semi-hard magnet and hence alters the total resultant magnetic flux.

Two pole pieces are attached to the ends of the magnet structure on both sides which helps to direct the flux in the desired direction. A current carrying coil is wound around the magnets to apply an external magnetic field. Coercivity of magnetic materials is a measure of how easily it can be magnetised and demagnetised [69]. The semi-hard magnet has a relatively low coercivity, in comparison with the hard magnet, and therefore its magnetic field can be altered relatively easily. Therefore, when current flows through the coil according to Ampere’s right-hand screw rule [69], an external magnetic field is formed, and the magnetisation of the semi-hard magnet will follow the same direction as the applied field. For example, in Figure 4.1(a) the magnetic field in the semi-hard magnet is from right to left which is the same as the external field generated by applied current. When the current is applied in the opposite direction in Figure 4.1(b) the semi-hard magnet forms a magnetic field from left to right. The hard magnet has a much higher coercivity and its magnetisation is not affected by the externally applied field as seen in Figure 4.1(a) and Figure 4.1(b). Therefore, the total magnetic flux formed when the current is discontinued can be contained internal to the magnetic structure as in Figure 4.1(a) or extend outside the magnetic structure as in Figure 4.1(b). When the flux is internal to the magnets the plunger remains unattracted and when the flux extends out of the magnets the plunger is attracted to facilitate the actuation.

### 4.1.1. Principles of Operation

The actuator needs to be designed precisely to meet certain conditions to successfully facilitate the actuation. The magnetic flux $\phi$ that passes through a cross sectional area $A$ of a material with magnetic flux density $B$, is defined by (4.1) [69].
\[ \Phi = B A \quad (4.1) \]

The magnetic flux in each material should be the same to ensure they cancel out any external field in the instance outlined in Figure 4.1 (a). If both magnets have the same dimensions the following conditions must be met:

- Hard and semi-hard magnetic materials must have similar residual magnetism
- Hard magnet must have significantly higher coercivity than semi-hard magnet
- Applied magnetic field must exceed the coercivity of semi-hard magnet but not the coercivity of hard magnet

The residual flux density \( B_r \), also known as remanence, is due to the magnetisation that remains in a ferromagnetic material once the applied external magnetic field is removed [69]. In this case, once the current through the coil is discontinued both materials rely on their residual magnetism to hold their magnetic fields. If \( B_r \) for both materials are similar or nearly similar, and both magnets have the same cross-sectional area, the magnitude of the flux generated in each material should be the same, according (4.1). If both materials generate the same amount of flux, when the materials are magnetised in opposite directions as in Figure 4.1(a), the flux lines will be contained within the magnetic structure to ensure that magnetic field remains internal. When the materials are magnetised in the same direction as in Figure 4.1(b), the flux lines add up to create an external magnetic field strong enough to attract the plunger.

Coercivity or the coercive force is the intensity of the magnetic field that needs to be applied to magnetise a ferromagnetic material by driving it to saturation [69]. The higher the coercivity \( H_c \), the harder it is to magnetise and demagnetise the material. The significance of \( H_c \) and \( B_r \) of a ferromagnetic material are reflected in the nature of its hysteresis loop [69]. Ferromagnetic materials with high coercivity are hard magnetic materials and make permanent magnets [69]. Materials with low coercivity are soft magnetic materials such as ferrite and certain types of alloys [69]. Materials with coercivities between soft and hard magnetic materials are known as semi-hard magnets [69], which mostly includes alloys. The selected materials should also have significantly
different coercivities to ensure that the hard magnet is not affected by the applied magnetic field, while the semi-hard magnet is easily magnetised.

It is very important to maintain the applied magnetic field \( H_{ap} \) at a value which will only impact the semi-hard magnet and not the hard magnet. This can be directly controlled by the coil current \( I \) as in (4.2), where \( N \) is the number of turns and \( h \) is the length of the coil [69].

\[
I \geq \frac{H_{ap} h}{N} \quad (4.2)
\]

Figure 4.2 illustrates the relationship between \( B \), \( H_c \) and the applied field \( H_{ap} \), for hard, semi-hard and soft magnetic materials.

\[
\begin{align*}
B_{r_{\text{soft}}} & \quad B_{r_{\text{semi hard}}} & \quad B_{r_{\text{hard}}} \\
H_{c_{\text{hard}}} & \quad H_{c_{\text{semi hard}}} & \quad H_{c_{\text{soft}}}
\end{align*}
\]

**Figure 4.2: Magnetisation curves for soft, semi-hard and hard magnetic materials**

### 4.1.2. Material Selection

The first task was selecting materials meeting the aforementioned criteria for successful actuation. Rare earth magnets such as Neodymium Iron Boron (NdFeB) and Samarium Cobalt (SmCo) with coercivities over 100 kA/m are hard magnets. Semi-hard magnets,
such as Aluminium Nickel Cobalt (AlNiCo), have lower coercivities than hard magnets usually in the range of 1-100 kA/m. Soft magnets have coercivities ranging between 0.1-1 kA/m. Several semi-hard and hard magnetic materials were studied for their coercivity and remanence, as depicted in Table 4.1 [75-77]. It can be seen that NdFeB and AlNiCo 5 have similar residual magnetism values at 1.12-1.41T and 1.1-1.35T respectively [78, 79]. These materials have vastly different coercivities at 800-2465 kA/m for NdFeB and 48-58 kA/m for AlNiCo [78-80]. Therefore, NdFeB was chosen as the hard magnet and AlNiCo 5 was chosen as the semi-hard magnet.

Table 4.1: Coercivity and remanence of magnetic materials

<table>
<thead>
<tr>
<th>Material</th>
<th>Coercivity (kA/m)</th>
<th>Remanence (T)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Hard Magnets</td>
<td></td>
<td></td>
</tr>
<tr>
<td>SmCo&lt;sub&gt;5&lt;/sub&gt;</td>
<td>1200-1800</td>
<td>0.9-0.95</td>
</tr>
<tr>
<td>Nd&lt;sub&gt;2&lt;/sub&gt;Fe&lt;sub&gt;14&lt;/sub&gt;B</td>
<td>800-2465</td>
<td>1.12-1.41</td>
</tr>
<tr>
<td>Sm&lt;sub&gt;2&lt;/sub&gt;Co&lt;sub&gt;17&lt;/sub&gt;</td>
<td>620-680</td>
<td>1-1.2</td>
</tr>
<tr>
<td>Ce(CuCo)&lt;sub&gt;5&lt;/sub&gt;</td>
<td>294</td>
<td>0.7</td>
</tr>
<tr>
<td>SrFe</td>
<td>230-275</td>
<td>0.38-0.4</td>
</tr>
<tr>
<td>BaFe&lt;sub&gt;12&lt;/sub&gt;O&lt;sub&gt;19&lt;/sub&gt;</td>
<td>175</td>
<td>0.24-0.36</td>
</tr>
<tr>
<td>Hard Ferrite</td>
<td>148-280</td>
<td>0.23-0.42</td>
</tr>
<tr>
<td>AlNiCo 8</td>
<td>110-150</td>
<td>0.72-0.9</td>
</tr>
<tr>
<td>Semi-hard Magnets</td>
<td></td>
<td></td>
</tr>
<tr>
<td>AlNiCo 5</td>
<td>48-58</td>
<td>1.1-1.35</td>
</tr>
<tr>
<td>AlComax 2-8</td>
<td>46-62</td>
<td>1.15-1.37</td>
</tr>
<tr>
<td>Soft Magnets</td>
<td></td>
<td></td>
</tr>
<tr>
<td>MnBi</td>
<td>0.9</td>
<td>0.48</td>
</tr>
<tr>
<td>Iron</td>
<td>0.08-0.16</td>
<td>0.77</td>
</tr>
<tr>
<td>FeNi</td>
<td>0.0008-0.08</td>
<td>0.04-6</td>
</tr>
</tbody>
</table>

Figure 4.3 illustrates the key parameters of the two chosen materials with their hysteresis loops plotted on the same graph.
Iron was chosen for the pole pieces, due to its ferromagnetic properties combined with the soft magnetic behaviour. In this context the pole pieces will always be subject to the magnetic fields of the NdFeB and AlNiCo magnets. Therefore, the magnetic field through iron is always dependent on these. The very low coercivity of iron at 0.08-0.16kA/m, makes its magnetisation and demagnetisation particularly easy. And its remanence is 0.77T which is relatively lower than NdFeB and AlNiCo and will have little or no impact in altering the total magnetic flux.

4.2. Electromagnetic Modelling and Analysis

The key magnetic parameters of the actuator were determined based on the physical laws governing it, while key electrical parameters were also modelled using circuit theory combined with electromagnetics.

4.2.1. Magnetic Circuit Modelling

The important parameters of the actuator were modelled mathematically using magnetic circuits shown in Figure 4.4. Each magnet is represented by the magnetomotive force it
generates and its reluctance, denoted as \( \text{MMF}_{NdFeB} \), \( \mathcal{R}_{NdFeB} \) for NdFeB and \( \text{MMF}_{AlNiCo} \), \( \mathcal{R}_{AlNiCo} \) for AlNiCo. The pole pieces attached to each side and the plunger are constructed of iron and represented by their reluctance \( \mathcal{R}_{\text{iron}} \). The reluctance of the airgap between plunger and magnet structure is denoted by \( \mathcal{R}_{\text{gap}} \). The magnetic materials are the combined to give \( \mathcal{R}_{\text{eq}} \) and \( \text{MMF}_{\text{eq}} \), while \( \mathcal{R}_{\text{iron1}} \) is negligibly small in comparison with the permanent magnets and therefore the original circuit is then simplified to its equivalent circuit model shown.

According to Hopkinson’s law [81] the magnetomotive force (MMF) for the two materials are as shown in equations (4.3) and (4.4), where \( \Phi \) is the flux of each material and \( \mathcal{R} \) is the reluctance.

\[
\begin{align*}
\text{MMF}_{NdFeB} & = \Phi_{NdFeB} \mathcal{R}_{NdFeB} \quad (4.3) \\
\text{MMF}_{AlNiCo} & = \Phi_{AlNiCo} \mathcal{R}_{AlNiCo} \quad (4.4)
\end{align*}
\]

The equivalent reluctance and MMF are calculated by equation (4.5) and (4.6). The reluctance is analogous to resistance in electric circuits and the parallel combination of the reluctances are calculated to quantify the equivalent reluctance \( \mathcal{R}_{\text{eq}} \). The flux generated by each magnet combines to create a strong magnetic field or a weaker magnetic field and hence a low reluctance path. The total flux generated is then multiplied by the equivalent reluctance to calculate the equivalent MMF. Depending on the direction of the current the orientation of \( \text{MMF}_{AlNiCo} \) is altered, a strong magnetic field or low reluctance path is created through the plunger to attract or release it.
\[ R_{eq} = \frac{R_{AlNiCo} \cdot R_{NdFeB}}{R_{AlNiCo} + R_{NdFeB}} \]  

(4.5)

\[ MMF_{eq} = \left( \frac{R_{AlNiCo} \cdot R_{NdFeB}}{R_{AlNiCo} + R_{NdFeB}} \right) (\Phi_{NdFeB} \pm \Phi_{AlNiCo}) \]  

(4.6)

Several views of the magnetic structure are shown in Figure 4.5 to assist the mathematical calculations that follow.

\[ B = \frac{B_r}{2} \left( \frac{h}{\sqrt{r^2 + h^2}} \right) \]  

(4.7)

\[ \Phi_{total} = \frac{\pi h r^2}{2 \sqrt{r^2 + h^2}} \left( B_{r,NdFeB} \pm B_{r,AlNiCo} \right) \]  

(4.8)

The total flux adds up when the magnets are magnetised in the same direction. When the magnets are magnetised in opposite directions, the difference between residual magnetisms quantify the total flux, which in this case is between 0-0.08 T ≈ 0, resulting in no flux in the external region.

When the residual magnetism values sum up to generate an external field, a force \( F_{gap} \) is generated in each air gap to attract the plunger, given by the formula derived in (4.9) [82].

\[ B_{gap} = \frac{\Phi_{total}}{ab} \rightarrow F_{gap} = \frac{(B_{gap})^2 ab}{2 \mu_0} \]

\[ F_{gap} = \frac{\left( \pi h r^2 \left( B_{r,NdFeB} \pm B_{r,AlNiCo} \right) \right)^2}{8 \mu_0 ab (r^2 + h^2)} \]  

(4.9)
As there are two identical airgaps on either side, the total force acting on the plunger is doubled, and it needs to exceed the weight of the plunger in order to attract it against gravity. Therefore, the maximum mass $m$ that can be pulled vertically by the EP magnet is given by (4.10) where $g$ is the gravitational acceleration.

$$m \leq \frac{[\pi h (r^2)(B_{rNdFeB} \pm B_{rAlNiCo})]^2}{8\mu_0 ab g (r^2 + h^2)} \quad (4.10)$$

### 4.2.2. Electric Circuit Modelling

While the EP magnet operates in the aforementioned stable states, the magnetisation current is delivered during a short transient period to change the polarization of the AlNiCo magnet. During this stage, the key priority is for the applied field to exceed the coercivity of 48kA/m for AlNiCo [79]. Since the actuation is infrequent, the losses in the AlNiCo core is ignored and for simplicity, the coil is treated as an air core solenoid which creates a magnetic field as a result of the current through it. To ensure strong magnetisation the applied field $H_{ap}$ should be at least three times the minimum field required [80], which demands a field of 150kA/m approximately. The current required to alter the magnetic field $I$ can be estimated by equation (4.11), where $N$ is the number of turns and $x$ is the magnetic path length [69]. When the plunger is attracted to the magnets, the magnetic path length $x$ is relatively longer at 11.2mm, with the added length of the pole pieces. However, in calculating the required current to magnetise AlNiCo in the opposite direction, the AlNiCo and coil combination is treated as solenoid of length $l=3.2$mm, with $N=200$ turns and the required H-field was determined from the hysteresis curve. The current required for an applied H-field of 150kA/m worked out to be 2.4A, as in the previous scenario.

$$I \geq \frac{H_{ap} x}{N} \quad (4.11)$$

The following first order DC transient circuit is the equivalent circuit model for the solenoid coil in Figure 4.6, which can be used to calculate the required voltage and pulse duration.
The current through the coil becomes a function of time due to the transient response of the inductor as shown in equation (4.12).

\[ I(t) = \frac{V}{R_{\text{coil}}} \left( 1 - e^{-\frac{tR_{\text{coil}}}{L_{\text{coil}}}} \right) \]  \hspace{1cm} (4.12)

The coil current should be able to supply the required field of 150 kA/m, with a supply voltage \( V \) given by equation (4.13).

\[ V = \frac{H_{\text{ap}} h R_{\text{coil}}}{N} \]  \hspace{1cm} (4.13)

Therefore, the time \( t \) taken to develop the desired current can be determined using the formula (4.14).

\[ t = -\left( \frac{R_{\text{coil}}}{L_{\text{coil}}} \right) \ln \left[ \frac{VN - hR_{\text{coil}}H_{\text{ap}}}{V N} \right] \]  \hspace{1cm} (4.14)

The series resistance and inductance of the copper coil can be calculated using equations (4.15) and (4.16), where \( L \) is the total length of the copper wire and \( \rho_{\text{copper}} \) is the resistivity of copper.

\[ L_{\text{coil}} = \frac{2\pi \mu_0 (N r)^2}{h} \]  \hspace{1cm} (4.15)

\[ R_{\text{coil}} = \frac{4 l \rho_{\text{copper}}}{\pi w^2} \]  \hspace{1cm} (4.16)

It was derived that the exact length of the wire \( L \) can be calculated using equations (4.17) and (4.18), based on the number of turns \( N \), radius of AlNiCo magnet \( r \), length of the solenoid \( h \) and the diameter of the coil wire \( w \).
\[ l = \frac{h}{w} \left[ 2nr(\pi + 2) + \sum_{i=1}^{n} 2\pi wi \right] \quad (4.17) \]

\[ n = \frac{Nw}{h} \quad (4.18) \]

For simplification purposes, it was estimated that the average turn length to be the length of the centre-most layer \( n/2 \). The total length of the coil is then given by equation (4.19).

\[ l = \frac{N}{2h} \left[ 4rh(2 + \pi) + \pi N w^2 \right] \quad (4.19) \]

The saturation current \( I_{\text{sat}} \) can be calculated using equation (4.20), where \( B_{\text{sat}} \) is the saturation flux density for AlNiCo. If \( t \) is the time taken for the coil current to rise to \( I_{\text{sat}} \) the energy \( E \) required to switch the direction of magnetisation can be estimated using equation (4.21).

\[ B_{\text{sat}} = \frac{L I_{\text{sat}}}{\pi r^2 N} \quad (4.20) \]

\[ E = V I_{\text{sat}} t \quad (4.21) \]

### 4.2.3. Magnetic Simulation Analysis

Magnetic modelling of the actuator was done using JMAG Designer, to analyse the behaviour of the flux patterns under different current conditions. The duration of the current pulse and the magnitude of the current were varied to analyse different conditions for the chosen dimensions. A current of 3.8 A pulsed through the coil for 200\( \mu \)s in either direction, and the 3-dimensional view of both scenarios are shown in Figure 4.7. In Figure 4.7(a) the current magnetises the AlNiCo in opposite direction to NdFeB and the flux is cancelled out with no external field present. In Figure 4.7(b) both magnets are magnetised in same direction to generate a total external flux. The results shown are generated once the coil current has been discontinued and the magnets are governed by their residual magnetism \( B_r \).
Figure 4.7: 3D view of FEMM simulation of external magnetic field generation

The top views of both scenarios are shown in Figure 4.8, which clearly illustrates the opposing flux lines in Figure 4.8 (a) and the flux lines in the same direction in Figure 4.8 (b). Figure 4.9 shows the magnetic flux patterns through a cross section of the AlNiCo magnet in both instances. Simulations indicated that the magnetic actuation concept is valid, and the external magnetic field can be turned on and off by pulsating a current through the actuator coil in either direction.

Figure 4.8: Top view of FEMM simulation
4.3. Actuation Control Design and Simulation

4.3.1. Magnetic Excitation Circuit Design

The actuator control is achieved in conjunction with the main ABL control platform, which delivers a wireless signal to actuate the valve. The microcontroller of the ABL platform commands a change in the IPT switching frequency to a predetermined value. In this instant the switching frequency is changed to 1 MHz and this change is identified by the pickup circuit inside the actuator. Once the change in frequency is detected by the secondary circuit steps are taken to actuate the microvalve. The actuation is controlled by two main circuits, the main control circuit and the position detection system. Details of the wireless IPT power supply are discussed in Chapter 5, and control signal delivery and detection is discussed in Chapter 6.
4.3.1.1. Main Control Circuit

Each part of the actuator is interconnected and Figure 4.10 shows the main components linked to the actuator control circuit.

Once the actuation command is received from the ABL platform it is picked up by the IPT coil and the control signal detection circuit generates a command to drive the actuator. This command turns on a switch between the supercapacitor energy buffer and the actuation circuit to enable power flow to the actuation circuit. This initiates a position detection command to the position feedback circuit to determine the next position of the actuator. The position feedback circuit drives one of its two output pins high depending on whether the actuator needs opening or closing. The logic circuit that follows generates a very short trigger pulse for one of the two timers. The timers are configured to operate in monostable mode and based on the values of the $R_3/C_3$ and $R_4/C_4$ combinations a precisely timed pulse is generated by the relevant timer for the H-bridge circuit. This micro-second range pulse determines the duration of the magnetisation current in the desired direction through the actuator. According to the principles discussed in Chapter 4, the change in magnetisation will attract or release the plunger to toggle the state of the valve. An Infineon TLE5206 H-bridge capable of handling up to 6A is used, due to the
high current required to switch the magnetisation. The supply voltage for the H-bridge can be varied between 7-20V, depending on the current required to alter magnetic field of AlNiCo.

### 4.3.1.2. Actuator Position Detection

Several options were considered to determine the present position of the actuator, such as the use of a micro-miniature tactile switch, the use of infra-red (IR) sensors and photo sensors. As shown in Figure 4. 11, one of the options were to use an optical signal on one side while an optical sensor was placed on the opposite side. If the plunger was attracted to the magnet structure the optical signal would be blocked and a low signal would be present at the output indicating the valve is open. If the plunger is repelled the optical signal would reach the sensor, generating a high output to indicate the valve is closed.

![Actuator position sense circuit using optical signal](image)

*Figure 4. 11: Actuator position sense circuit using optical signal*

An IR LED was the best choice as the optical signal source and several sensing options were investigated are outlined in Figure 4. 12. The LED and phototransistor option shown in Figure 4. 12 (a) generates a high output signal when the valve is closed, and the LED light is incident on the phototransistor. A type KPA 3010 was chosen as the phototransistor mainly due to its miniature dimensions. The photoresistor based sensing technique shown in Figure 4. 12 (b) works in the same manner, to generate a high output.
when the valve is closed. In Figure 4.12 (c) a colour sensing photodiode, KPS 5130, is used to sense a particular coloured light. In this case a red LED was considered as the light source to produce an output current to generate a high output signal. In these optical solutions, the presence of ambient light was problematic and required the sensor to be completely isolated optically, requiring a very tight mechanical design. The photodiode option was less sensitive to ambient light as it was designed for a specific frequency of the light.

![Optical detection techniques](image)

Figure 4.12: Optical detection techniques

A more feasible solution was found using a micro tactile switch to generate a short pulse, when the valve is opened or closed. Figure 4.13 illustrates the sensing mechanism in detail, where tactile switched as placed under each pole piece and will close the circuit when the valve is opened, and the plunger is attracted to the magnets.
Due to the real estate constraints within the actuator, the tactile switch needed to be chosen meticulously. Several switches, as outlines Table 4. 2 were investigated for feasibility.

<table>
<thead>
<tr>
<th>Tact Switch Model</th>
<th>Dimensions</th>
<th>Actuator Height</th>
<th>Operating Force</th>
</tr>
</thead>
<tbody>
<tr>
<td>E switch: TL3315NF100Q</td>
<td>4.5 x 4.5 mm</td>
<td>0.55 mm</td>
<td>100 gf</td>
</tr>
<tr>
<td>Panasonic: EVP-BB4A9B000</td>
<td>2.6 x 1.6 mm</td>
<td>0.53 mm</td>
<td>240 gf</td>
</tr>
<tr>
<td>Citizen: LS70C2-T</td>
<td>2.8 x 2.3 mm</td>
<td>0.35 mm</td>
<td>160 gf</td>
</tr>
<tr>
<td>Citizen: LS95C2D-T</td>
<td>2.8 x 1.75 mm</td>
<td>0.55 mm</td>
<td>160 gf</td>
</tr>
<tr>
<td>Citizen: LS75C2D-T</td>
<td>2.8 x 1.9 mm</td>
<td>0.5 mm</td>
<td>160 gf</td>
</tr>
<tr>
<td>Panasonic: EVQ-P6PB35</td>
<td>4.1 x 4.1 mm</td>
<td>0.35 mm</td>
<td>240 gf</td>
</tr>
</tbody>
</table>

The Citizen LS75C2D-T switch was chosen as it offered a good compromise between operating force, dimensions and actuator height. It fits well with only a small portion of its width outside each pole piece with dimensions of 1.3 x 3.2 mm. Figure 4. 14 shows the implemented position feedback circuit. When the valve is in the open position, the switch is pressed and both comparators are supplied with a high 3V signal. This signal is fed to the inverting input of the first comparator and the non-inverting input of the second comparator. Both comparators have a 1 V reference supplied by the voltage divider, which gives a high output 1 and a low output 2, when the switch is pressed. When the valve is closed, and the switch is uncompressed output 1 is low and output 2
is high. These signals are then supplied to the actuator control circuit outlined in Figure 4. 10.

![Actuator control circuit diagram](image)

*Figure 4. 14: Implemented position feedback circuit*

This circuit worked effectively in detecting the present position of the actuator, which allowed the logic circuit to determine the current flow direction through the EP magnet coil to toggle the valve.

### 4.3.2. Actuator Control Circuit Simulation

The behaviour of the actuator control circuit was simulated in LTSpice to ensure accurate actuation, and the comparator was fed a short pulse to imitate the pulse generated by the micro-switch in the position detection circuit. The output of the comparator, the trigger pulse input generated by the logic circuit and the actuation current pulse delivered to the EP magnet coil are shown below in Figure 4. 15.
It can be seen that the trigger pulse in this instant is generated only when the comparator output is on a rising edge, which means the position feedback circuit is transitioning from high to low. This translates as the valve being in the open state, and the current is required to drive the coil to close the valve. The second timer behaves in the same manner, with the only difference being the comparator generates a high output when its input goes from low to high, generating a current pulse to the coil to open the valve.

4.4. Actuator Implementation

A proof of concept actuator was implemented using the chosen materials, with the dimensions outlined in Table 4. 3, which were the same dimensions used in the FEMM simulations.

Table 4. 3: Dimensions of magnetic structure
Table 4.1: Dimensions of the Bi-Stable Actuator

<table>
<thead>
<tr>
<th>Dimension</th>
<th>Length</th>
</tr>
</thead>
<tbody>
<tr>
<td>Radius of magnets</td>
<td>r</td>
</tr>
<tr>
<td>Length of magnets</td>
<td>h</td>
</tr>
<tr>
<td>Height of pole piece</td>
<td>x</td>
</tr>
<tr>
<td>Width of pole piece</td>
<td>a</td>
</tr>
<tr>
<td>Thickness of pole piece</td>
<td>b</td>
</tr>
<tr>
<td>Thickness of coil wire</td>
<td>w</td>
</tr>
</tbody>
</table>

Although the coil is only required to be wound around the AlNiCo magnet, due to practical difficulties it was wound around both magnets. Therefore, the applied field was maintained well under the coercivity levels of NdFeB to ensure its magnetisation remained unchanged. The implemented actuator is shown in Figure 4.16.

![ Implemented Actuator ]

_Figure 4.16: Implemented actuator_

The implemented actuator control and position feedback circuit is shown in Figure 4.17. A final prototype of the control circuit was created using surface mount ICs to miniaturise the system. However, it has a lot more allowance for optimal miniaturisation to fit in the valve enclosure, which is one of the future recommendations made in Chapter 8. The micro tact switch of the position feedback circuit was carefully attached to each actuator pole piece. However, in Figure 4.17 only one switch is attached to illustrate more clearly.
Preliminary experiments were conducted to establish the actuation concept, which proved successful. The pulses used to drive the actuator were recorded on the oscilloscope as shown in Figure 4.18.
Iron powder was used to observe the formation of the magnetic flux patterns of the two instances when the magnets have the same magnetisation and opposite magnetisation as shown in Figure 4. 19.

A strong magnetic flux pattern and attraction force was experienced when the magnets were polarised in the same directions as seen in Figure 4. 19 (a). The flux patterns were much weaker when the magnets had opposite polarisations as seen in Figure 4. 19 (b) with nearly zero attraction force.

4.5. Experimental Results and Analysis

4.5.1. Number of Turns versus Energy Consumption

The first experiment was to determine the optimum number of turns for the magnetic structure, where the minimum energy consumption was observed. The number of turns was varied in multiples of 10, from 100 to 300. The theoretical minimum and ideal current
required to achieve an applied field of 48 kA/m and 150 kA/m were calculated for each instance using equation (4.2). During experiments the deflection distance, the distance between the plunger and magnetic structure, was maintained at 2mm and the minimum supply voltage to attract the plunger was recorded. The minimum supply voltage corresponds to when the plunger was attracted with a bare minimum attraction force. The resistance of the coil was measured against the number of turns and used to calculate the magnetisation current in the coil for each instance. This data was used to calculate the minimum current to magnetise, which is governed by equation (4.2). The minimum current required to generate an applied field of 48 kA/m was calculated using equation (4.2). Results for magnetisation current and the applied magnetic field are plotted against the number of turns in Figure 4.20, along with theoretical predictions for minimum and ideal currents.

The current measured at the borderline magnetisation point for AlNiCo, which is when the plunger was barely attracted to the magnet, falls well between the theoretical minimum and ideal currents for each instance. The applied magnetic field strength at the borderline point is plotted, which shows fluctuations when the number of turns is under 200, with a minimum at about 120 turns. Although a minimum magnetic field is observed at 120 turns, a trend is seen at 200 turns where the required magnetic field gradually increases beyond 200 turns. When the coil has two hundred turns a magnetic field of

![Figure 4.20: Magnetisation current & applied magnetic field against number of turns](image-url)
71.13 kA/m and a current of 1.14A was applied to the coil. The reason for these fluctuations when the number of turns is less than 200 is unknown and will need to be investigated in future.

Further experiments were conducted to observe the magnetisation current and supply voltage requirement to barely attract the plunger at 2mm, with a fixed pulse width of 106.5 µs. It confirmed that supply voltage and current requirements were optimized at 200 turns, as seen in Figure 4.21. The magnetisation current dropped significantly for a coil with 200 turns or more. But as the number of turns increased beyond 200 the required supply voltage increased gradually while the magnetisation current remained reasonably constant. This was mainly due to the increase in the coil resistance as the length of wire increased, according to equation (4.16).

![Figure 4.21: Magnetisation current, supply and coil voltage against number of turns](image)

The power and energy consumption of actuator were calculated and plotted against the number of turns in Figure 4.22. Results indicate that the power and energy consumption of the actuator is utilised mainly for the magnetisation of the AlNiCo magnet, while the remaining circuitry consumes constant and minimal power and energy. Power and energy consumption are both minimal at 200 turns, making this the optimum number of turns for the designed actuator. It can also be noted that the minimum energy...
consumption at 200 turns is less than 1mJ, which is the most attractive feature of this actuator.

![Figure 4.22: Input/consumed power and energy against number of turns](image)

**4.5.2. Pulse Duration versus Energy Consumption**

Due to the difficulty in changing the number of turns when conducting repetitive experiments, the number of turns was maintained at 200, according to the optimal value indicated by experiments and theory. Follow up experiments were carried out by changing the input voltage to control the coil current. The duration of the current pulse was varied, using variable capacitor control, to identify the minimum time the current should be applied for, at each voltage level to actuate valve. The energy consumption for each instance was calculated and plotted against the supply voltage in Figure 4.23. The energy requirement is estimated based on the same pulse width. The applied voltage was calculated theoretically using equation (4.21), and is plotted against the measured values in Figure 4.23. While the actuation energy patterns for turning the external field on and off are slightly different, it is seen that the theoretically calculated energy estimate, although slightly lower, aligns very closely with the measured energy consumption.
The energy consumption and pulse width for turning the external field on and off, was plotted against the supply voltage in Figure 4.24. The minimum actuation pulse was as small as 37 µs and can be as large as 158 µs. This aligns well with the theoretical model which shows that the current in the actuator coil will only stabilize after 49 µs and a pulse longer than this should be required to ensure stable magnetisation of AlNiCo. Results show that the current pulse width decreases with an increasing voltage/current, resulting in an increase in the energy consumed. The energy consumption per actuation increases as the energy is proportional to the square of the voltage/current. Therefore, it was deemed feasible to supply a lower voltage for a longer time, ideally 12 V to minimise the energy consumption.
4.5.3. **Holding and Attraction Force of Actuator**

Next the physical performance parameters of the actuator such as holding force, attraction force and deflection range were investigated. The holding force is identified as the maximum force the actuator can exert on the plunger to hold it in the attracted position. The attraction force is defined as the maximum force the actuator can exert on the plunger when it is in not already attracted and is placed at a particular distance from the actuator. To investigate the holding and attraction forces of the actuator the supply voltage/current was varied with a known weight attached to the plunger. Figure 4.25 shows the plunger with additional weights attached to test its holding force capability.

![Figure 4.25: Actuator and plunger with additional weight attached](image)

When the external magnetic field was turned on the maximum weight the magnet could attract at a specified deflection distance was recorded against the supply voltage. Experiments were repeated at different pulse widths where the attraction force was calculated using the maximum weight attracted by the actuator for different magnetisation currents as seen in Figure 4.26. It was seen that attraction forces as large as 98 mN were observed. It was evident that higher magnetisation currents enabled higher attraction forces, while the pulse width had little effect on the attraction force.
While the plunger remained attracted to the magnet additional known weights were attached until the magnet could no longer hold the plunger, where gravity exceeded the attraction force of the actuator. This added weight was recorded against each supply voltage level, which was later used to quantify the holding force shown in Figure 4. 27. It was seen that longer pulses and higher currents, which deliver larger amounts of magnetisation energy, facilitated higher holding forces. The maximum holding force observed was 220 mN. This is well under the maximum force capability estimated using equation (4.10), which shows that the maximum force should be under 909 mN.
The magnetisation energy was calculated for each instance as the pulse width was varied for each experiment and the maximum attraction force at 1 mm and the maximum holding force were plotted in Figure 4.28. The holding force was always higher than the attraction force as the attraction force is in effect at a deflection range of 1 mm while the holding force comes into effect at the interface between the magnet structure and the plunger.

![Figure 4.28: Maximum holding force and attraction force against magnetisation energy](image)  

Repeated experiments were conducted to determine the relationship between the attraction force and deflection for different pulse widths. Deflection refers to the displacement of the plunger as a result of the actuator forces. Data gathered for a pulse width of 126 µs is plotted in Figure 4.29. It was seen that with deflections under 1 mm the attraction force stayed constant, while attraction force dropped beyond 1 mm. It was also seen that higher magnetisation currents allowed for higher attraction forces and deflections.
4.5.4. Deflection Range of Actuator

The supply voltage/current was varied, and the plunger was placed at known distances to be attracted. The maximum possible separation between the magnet structure and the plunger before attraction is no longer possible was recorded against the pulse voltage. This was taken as the maximum deflection, which is plotted against the magnetisation current in Figure 4.30. It was observed that with an increased magnetisation current, the deflection increases, as a stronger magnetic field is formed. Deflections less than 0.3 mm were not practically realisable as the weak magnetic field in the off state attracts the plunger. The maximum deflection observed was 2.5 mm when the current was at its highest value of 3.1 A.

Figure 4.29: Maximum attraction force against deflection

![Graph showing attraction force against deflection](image)
The pulse width was varied for different pulse voltage levels to observe the maximum deflection achievable, as shown in Figure 4.31. It was observed that the deflection increased with the pulse width, although it was constant for a certain range of pulses, before the deflection increases again.

The minimum deflection energy required for a particular deflection range was recorded and plotted in Figure 4.32.
4.6. Performance Summary

The experimental results can be summarized as seen in Table 4.4 which shows the best operating conditions for the actuator. The developed actuator is highly energy efficient and provides many attractive attributes such as fast actuation, large forces and deflection. The only significant drawback is the peak power requirement of 40 W, which too is met by a supercapacitor energy buffer recharged by IPT wireless power, which will be discussed in Chapter 5.

<table>
<thead>
<tr>
<th>Table 4.4: Summary of actuator performance</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of turns</td>
</tr>
<tr>
<td>Applied magnetic field</td>
</tr>
<tr>
<td>Coil current</td>
</tr>
<tr>
<td>Supply voltage</td>
</tr>
</tbody>
</table>
Unlike the previously described microvalves in section 2.4, the number of actuations is directly proportional to the total energy consumption of this actuator. Therefore, this actuator is highly energy efficient in comparison with the solenoid valve currently used in the ABL platform.
Due to the significantly lowered energy demand of the novel actuation mechanism, adding on wireless power became a lot more feasible. Possible wireless power supply technologies were investigated to identify the most feasible solution for this microactuator.

When enabling wireless power one major challenge was meeting the peak power requirement of 40 W during the 120 µs actuation period. Meeting this power requirement purely using a miniature wireless power supply is challenging. Therefore, an energy buffer was required to store energy onboard to be released during actuation. Energy buffer options were also explored and a supercapacitor based energy buffer was designed for this purpose.

### 5.1. Review of Wireless Power Supply Technologies

Wireless power transmission (WPT) systems are broadly categorised as non-radiative and radiative based on the power transfer distance, which can be short to medium range or long range. Near field power transfer is in the range within the wavelength (1 \( \lambda \)) and the power transfer is non-radiative [83]. In this region, the electric and magnetic fields are separate and power transfer can be done using either of these fields. Capacitive coupling uses the electric field for power transfer [84] while inductive coupling utilises the magnetic field [73]. Power is not transmitted if a receiving device is not present in
this power transfer range, which is dependent on the size and shape of the transmitting and receiving coils. In non-radiative WPTs the power transmitted decreases exponentially over distance and if the distance between transmitter and receiver is greater than the coil diameters power transfer becomes very inefficient [85].

When the power transfer distance is greater than one wavelength the electric and magnetic fields become perpendicular to each other to propagate electromagnetic waves such as radio waves, microwaves and light waves [83]. This energy is radiative, which means energy is transmitted irrespective of whether or not a receiver is present to absorb energy. The energy loss of these systems can be much greater, but allows for much greater power transfer distances. The omnidirectional radiation of power allows very little power to be transmitted in the desired direction of the receiver and therefore is not efficient over a very long power transfer range. Well-designed antennae are required to concentrate the energy beams to improve the efficiency. Microwave and laser power transfer techniques are considered radiative [83]. Figure 5.1 classifies the different WPT systems currently in existence.

![Figure 5.1: Classification of wireless power transfer systems](image)

A qualitative comparison of the different WPT strategies is summarised in Table 5.1.
<table>
<thead>
<tr>
<th>Power Transfer Method</th>
<th>Advantages</th>
<th>Disadvantages</th>
<th>Power Level</th>
<th>Power Transfer Range</th>
</tr>
</thead>
</table>
| **IPT**               | Magnetic field coupling | • Relatively mature technology  
• High power capability | • Cannot transfer power across metal objects  
• Precise coupling design requirements  
• Shielding of magnetic field is vital | mW - kW | Medium |
| **CPT**               | Electric field coupling | • Transfer power through metal barriers  
• Simple coupling structure – alignment requirements not as precise as IPT  
• Low EMI  
• Light weight & small volume | • More practical for low power applications  
• Hazardously high voltage requirement at high power levels  
• High electric field intensity leads to production of reactive ozone  
• Interaction of electric fields with most materials, including human body | mW - W | Short |
| **UPT**               | Acoustic wave propagation | • Ability to power through metal barriers | • Immature technology  
• Lack of spatial separation | mW - W | Medium |
| **Laser**             | Light wave | • Very narrow beam can be used to achieve larger distance  
• Compact size:  
• No RF interference  
• Easier access control | • Hazardous  
• Low efficiency in conversion between electricity and light  
• Potentially up to 100% losses by atmospheric absorption  
• Requires direct line of sight with the target | kW | Long |
| **Microwave**         | Electromagnetic wave propagation | • Long transmission distance | • Limited to line of sight propagation | kW | Long |
Due to the maturity and availability it was decided to use inductive power transfer (IPT) as the onboard WPT system for the actuator. As the actuator required continuous power in the milli-watt order IPT supplies can deliver power to the target application.

With the increasing popularity of wireless power, standards have been put in place to ensure all devices work well together. Three different governing bodies have set out separate standards. The Wireless Power Consortium (WPC) established in 2008, being the oldest standards body, has established the Qi standard at 200 kHz for inductive power transfer [90]. On the other hand, Power Matters Alliance (PMA) chaired by WiTricity uses the 6.78 MHz band [90]. The Alliance for Wireless Power (A4WP), backed by Samsung and Qualcomm, has created the Rezence standard which is the same as the WiTricity standard [90]. These two bodies have now combined to form the AirFuel Alliance and compete with the Qi standard [90].

5.2. Inductive Power Transfer for Microactuator

Inductive power transfer (IPT) is the oldest and most widely used WPT topology to date and is found in most of the aforementioned commercial applications. The concept is the same as a transformer where a transmitting coil is excited with an alternating current (AC) to generate an oscillating magnetic field according to Ampere’s law [69]. A receiving coil placed in its vicinity is magnetically coupled and a secondary current is induced according the Faraday’s law of induction [69]. This current can directly drive the load or be rectified to produce a direct current (DC) to drive a load.

A sub-class of IPT is resonant IPT, where the secondary pickup is tuned to resonate at the same frequency as the primary IPT transmitter. A compensation network is formed on both the primary and secondary coils to achieve tuning, making it more efficient at greater distances [73]. Resonant coupling achieves a high quality factor $Q$, defined by (5.1) and (5.2) for series and parallel tuned RLC circuits, and power is exchanged at a higher rate than the internal damping [73].
\[ Q = \frac{1}{R} \sqrt{\frac{L}{C}} \]  
(5.1)  
\[ Q = R \sqrt{\frac{C}{L}} \]  
(5.2)

Figure 5. 2 depicts the basic concept of IPT and resonant IPT.

Introducing wireless power to the ABL platform has many advantages over conventional wired power supplies [73, 91].

- Freedom of mechanical movement
- Safe operation in chemical and liquid environment
- Zero corrosion and mechanical wear and tear
- Reliability and Robustness
- Water and liquid proof
- Improved aesthetics
- Better image acquisition

With wired power, movement is only allowed for the length of the power cord. The loose coupling between the primary and secondary allows the microactuators to move within a reasonable distance without losing power. New wirelessly powered microfluidic components can be easily appended to the system, without needing to redesign the power connections. Wireless power facilitates freedom of mechanical movement within the ABL platform.
Wireless power eliminates exposed connectors which are potentially hazardous in a liquid environment. Safe and spark free operation is a major advantage in sensitive systems such as the ABL platform. The absence of wired connectors ensures no harbouring of chemical residue, dirt, dust, water, bacteria and other contaminants, and will maintain a clean, safe and healthy environment in the microfluidic system.

Breaking of wires due to vigorous movement, mechanical wear and tear over time, chemical erosion etc. will no longer be a problem with a wirelessly powered ABL platform. This makes a WPT system a lot more reliable and robust with little maintenance required.

Wireless power eliminates the need for a physical connector allowing devices to be completely sealed and hence waterproof.

Wireless power enhances convenience, versatility and aesthetics of the chaotic ABL platform. When using a camera to acquire high resolution images for post experimental processing it is vital that there are no loose wires hanging around, potentially obstructing the view and causing unwanted shadowing to compromise the image quality. Although removing the tubing is impossible, eliminating the power connections is much more viable. This significantly improves the aesthetics of the platform allowing for better image acquisition.

However, adapting wireless power in the ABL platform has a several drawbacks such as:

- Reduced efficiency
- Additional cost and complexity of circuitry
- Biological effects on human beings

While a wired power supply delivers all the power to the microfluidic components, wireless power supplies have approximately 85-90% efficiency. This is mainly due to the losses DC/AC and AC/DC conversion processes and the loose coupling between the primary and secondary circuits. A significant portion of research in wireless power is dedicated to improving the efficiency of these systems to make them more desirable.
The cost and complexity of setting up a WPT system is significantly higher than a conventional wired power system, due to the need for additional circuitry. It is a trade-off between cost and convenience. However, with cost efficient design, this can be mitigated and with the ABL platform being a low power system, the additional cost is not as significant as a high power IPT system.

The biological effects of wireless power transfer varies based on the topology and frequency. With the increasing popularity of WPT, much research has gone into investigating the biological effects of WPT on human beings. Research conducted by the University of Auckland [92] shows that very low frequency (VLF) electromagnetic fields, such as the 10 kHz IPT converters have no negative biological effects. This is possibly due to these frequencies being far higher than those naturally occurring in the human body. As these frequencies are well under the RF range heating in the human body is not expected.

The wireless power supply of the microactuator is targeted to operate in the frequency range of several MHz, which falls into the shortwave category on the RF spectrum. Research has shown that at these frequencies heating of highly conductive tissue in human bodies may occur, which is in fact used as a form of physiotherapy known as shortwave diathermy [93].

5.3. Wireless IPT Power Supply Design

The wireless power supply has 3 main components:

- Track and coil design
- Primary resonant converter design
- Secondary pick-up circuit design

The primary track serves as the power transmitter while the secondary pick-up coil operates as the receiving antenna. The magnetic coupling of the IPT power supply is
mainly dependent upon the track and coil design. FEMM software, such as JMAG, CST etc. are often used to design the magnetic coupling, to ensure higher efficiency levels.

The primary resonant converter converts a DC power input to a high frequency AC power output, which is then fed to the resonant tank. The time varying magnetic field is generated by this AC signal, which helps deliver power to the secondary circuit through magnetic induction. There are several converter topologies which deliver different features, and these were investigated to identify the best option for the ABL platform.

The secondary pickup circuit performs two main tasks, couple with the primary converter to induce a secondary current and rectify and condition the power to be delivered to the load.

### 5.3.1. IPT Setup

In IPT a 10 – 100 kHz, high frequency AC signal is generated on the primary side to generate an alternating magnetic field, which is necessarily an inductor $L_p$. This field is coupled loosely with the secondary pickup coil, another inductor $L_s$, where a secondary AC current is induced. This current is then rectified, conditioned and delivered to the load. The basic IPT setup is shown in Figure 5.3.

![Figure 5.3: Basic IPT setup in the microactuator](image)

The mutual magnetic coupling between the primary and secondary sides is dependent upon the core material, the number of turns and the geometry of the coils [73, 94]. Although the coupling coefficient between two coils is dependent on the self-inductance of each coil and the mutual inductance, IPT systems have a lumped track where only part
of this track is coupled with the secondary causing the effective coupling to be much lower than the calculated value [69]. The coupling of such systems is less than or around 1%.

5.3.2. Track and Coil Design

In typical IPT systems the magnetic coupling is less than 30%, which is a trade-off to make way for mechanical freedom. This is caused by the high leakage inductance of coils due to the large air gap between the transmitter and receiver [73]. Alignment of the primary and secondary coils will ensure optimal performance, but in the ABL platform, as with many other practical IPT applications, perfect alignment of coils is highly unlikely. Therefore, the magnetic design of the transmission coils is very important to maximise the power transfer capability of the system.

The primary track distributes the generated magnetic flux, which then couples with the secondary side. Many different track layouts have been designed for various purposes, while Figure 5. 4 illustrates some of the common primary track arrangements [95].

Parallel cables, shown in Figure 5. 4(a) are suitable for applications where the load moves along the track as in monorail trolleys etc. [96]. The partial parallel configuration, as shown in Figure 5. 4(b), is applied for multiple stationary loads placed at the points where the track is not twisted, detached from the power source as in road way studs [97]. The inductance and track current are high while the inductance is negligible in the twisted parts [73]. For applications where the load is local to the power supply, such as biomedical implants [98], portable consumer devices [99] etc., the lumped coil, shown in
Figure 5.4(c) is far more suitable. In the ABL platform a single turn coil is used, to be able to deliver power to the multiple microactuators, which was deemed feasible due to the simplicity in design and reasonably efficient coupling.

Since the primary track is lengthy it is not fully coupled with the pick-up, and its’ inductance $L$ can be expressed in Henry per meter, according to equation (5.3). $\mu$ is the permeability, $d$ is the cable diameter and $D$ is the spacing between the conductors.

$$L = \frac{\mu}{\pi} \left[ \frac{1}{4} + \ln \left( \frac{2D}{d} \right) \right] \quad (5.3)$$

To mitigate the increase in ESR at high frequency due to skin and proximity effects, the tracks are usually constructed using Litz wire.

Due to the lower coupling, the amount of power transferred to the secondary depends on the mutual inductance between the two coils, $M$. $M$ is dependent on the core material, number of turns, and geometry of the coils [73]. Although a thorough magnetic analysis of the system can quantify $M$ analytically, it can be quantified by measuring the induced open circuit voltage $V_{oc}$ and using equation (5.4). $\omega$ is the angular operating frequency and $I_p$ is the primary current.

$$M = \frac{V_{oc}}{j\omega I_p} \quad (5.4)$$

The degree of coupling between two coils is defined by the coupling coefficient, $k$, as shown in (5.5) where $L_p$ and $L_s$ are the inductances of the primary and secondary coils.

$$k = \frac{M}{\sqrt{L_p L_s}} \quad (5.5)$$

Most often the track is not a lumped coil and only part of its inductance is linked with the pickup. The coupling coefficient calculated according to (5.5) using the total primary inductance would be much lower than its actual value. The coupling factor, $k_f$, defined by (5.6) better quantifies the coupling between primary and secondary. $N_p$ and $N_s$ are the number of turns in primary and secondary inductors. $I_{sc}$ is the short circuit current measured for the pickup, which is expressed in equation (5.7).
\[ k_f = \frac{N_s I_{sc}}{N_p I_p} = \frac{n_s I_{sc}}{I_p} \quad (5.6) \]

\[ I_{sc} = \frac{V_{oc}}{j\omega L_s} \quad (5.7) \]

Further simplification of the coupling factors is given by (5.8).

\[ k_f = \frac{n_s M}{L_s} \quad (5.8) \]

As the actuator does not require a constant power supply momentary disconnection of the IPT primary track is allowable. Therefore, a more flexible, non-restrictive single wire track design is utilised after magnetic analysis and experimental verifications.

The IPT secondary coil is usually a circular or rounded rectangular coil as shown in Figure 5.5, although it can be customised for specific applications. The other main factor determining the magnetic coupling is the area covered by the winding, which is severely limited in this application [100]. Research has shown circular coils provide the highest degree of coupling for a fixed inductor area as opposed to square and rectangular coils, as the magnetic field tends to distort at the corners of these inductors [100]. Other factors such as the diameter of the coils and the turns ratio of the IPT system has no influence on the coupling co-efficient \( k \).

![Figure 5.5: Secondary pick-up coils](image)

As the magnetic permeability plays a significant role, materials such as ferrite are introduced to increase the relative permeability. Ferrite is the most popular solution for the inductor core material due to its high magnetic permeability and high electrical resistivity [69]. This significantly reduces undesirable eddy currents and lowers losses at
high frequencies. A carefully designed ferrite structure can also guide the flux lines as desired to minimise leakage flux.

Increasing the effective area of the pick-up coil increases the coupling, which in this case is constrained by the dimensions of the microvalve. Due to the difficulties associated with constructing an efficient pickup coil it was decided to utilise well-constructed commercially available IPT coils by Würth Elektronik, as shown in Figure 5. 6. Experiments were carried out using the three pick up coil types to identify the best option, which was backed by magnetic simulations.

![Figure 5. 6: Commercially available charging coils tested on the microactuator](image)

The performance of the secondary pick up us determined by the induced open circuit voltage, \( V_{oc} \) and the short circuit current \( I_{sc} \). At frequency \( \omega \), with a primary current \( I_p \) and a mutual inductance of \( M \), \( V_{oc} \) is defined by equation (5. 9) while \( I_{sc} \) is defined by (5. 7).

\[
V_{oc} = j\omega MI_p \tag{5. 9}
\]

The short circuit current is given by (5. 7) as well as (5. 10).

\[
I_{sc} = \frac{MI_p}{L_s} \tag{5. 10}
\]

5.3.3. Tuning Circuit Design

Several key factors need to be considered when designing an IPT system for a specific operating frequency, such as compensation circuit options and the Q-factor. For a particular IPT topology, once the operating frequency is chosen, the compensation circuit
ensures resonant operation while the $Q$ factor is determined by the component values of the tuning circuit. The $Q$ factor has a significant influence on the efficiency of the power transfer circuit.

### 5.3.2.1. Operating Frequency Selection

IPT systems can operate in a vast range of frequencies varying from a few kHz to a several hundred MHz. The primary resonant converter is fed with a DC input, which is then converted to an AC output using the primary converter. This can be done using linear amplifiers or switch mode converters, the latter being the more popular choice [73]. The switching frequency of the converter is taken as the operating frequency of the IPT system which governs many parameters such as power loss, the magnetic coupling and transmission efficiency.

Increasing the operating frequency reduces the required primary current $I_p$ for a fixed power level. The coil current determines the amount of magnetic flux generated, which governs the core losses. By reducing the magnetic flux required for a specific power level, core losses can be minimized, which can be achieved by increasing the operating frequency [100].

Power losses in the inductors occur in three forms, conduction losses $P_w$, core losses $P_c$ and eddy current losses $P_s$ [100, 101]. $P_w$ and $P_s$ are proportional to the square of the winding current $I_p$ and the operating frequency $f_0$ as shown in the equations (5.11) and (5.12) below [100].

$$\begin{align*}
P_w &\propto I_p^2 \cdot \left(\frac{f_0}{\sqrt{f_0}}\right) & n < 2, low f_0 \\
P_s &\propto I_p^2 \cdot \sqrt{f_0} & high f_0
\end{align*}$$

Conduction losses occur due to the skin and the proximity effects. However, due to the sinusoidal excitation, the losses per unit volume of the core $P_c$ is proportional to the frequency of operation $f_0$ [100]. As skin and proximity effects are minimised by using litz wire in the primary conductor a higher frequency can be used to minimise the other
losses. Therefore, increasing the operating frequency up to an optimum value decreases losses.

Operating frequencies vary depending on the power level and real estate available for each application. Figure 5.7 shows a comparison of transmission frequencies and associated power levels for different IPT applications [100]. It can be seen that the operating frequency is increased as the power level of the application reduces.

![Figure 5.7: Power level against operating frequency of typical IPT applications [100]](image)

With fast-switching semiconductors power transfer frequencies in the MHz range are attainable [100]. Increasing the operating frequency also leads to smaller coil sizes for a fixed air gap, as the reduced magnetic coupling $k$ can be compensated for by the higher coil quality factor $Q$. The operating frequency selection is also influenced by the leakage flux emission near the coils [100]. However increasing the operating frequency has limitations, such as

Initially the operating frequency was determined at 6.78 MHz, to align with the Rezence standard, but the highest stable frequency practically achievable for the different converter types were 3.5 MHz and 800 kHz.
5.3.2.2. Compensation Network

While the switching frequency of the converter determines the operating frequency, a compensation network can help achieve resonance in the converter circuit for higher efficiency. The IPT primary track can drive the pick-up directly, without the need for additional reactive components. But in practical circuits additional capacitors and inductors are used to compensate for the large track inductance and achieve resonance [73].

Due to the large number of turns and the length of the track, the primary inductance may be too large to be driven by the power supply. Addition of a capacitor either in series, parallel or a composite arrangement as in Figure 5.8, compensates for this and makes a short track appear longer [73]. A compensation capacitor reduces the reactive power delivered by the supply [100], resulting in a highly selective band-pass characteristic of the input impedance of the circuit.

![Diagram of compensation circuits]

*Figure 5.8: Track compensation circuits*

The resonant frequency of the primary side is determined by (5.13):

\[
 f_0 = \frac{1}{2\pi \sqrt{L_p C_p}} \quad \text{and} \quad \omega_0 = \frac{1}{\sqrt{L_p C_p}} \quad (5.13)
\]

To achieve the resonant frequency of 800 kHz, the single wire primary inductor was measured at 1.5 µH and needed to be combined with a parallel capacitor of 26.38 nF, according to (5.13). However due to the stray inductances and capacitances in the circuit the capacitor value was adjusted accordingly.

Although tuning the pick-up coil to the operating frequency of the track is not essential for the operation of the IPT powered ABL platform, when the primary and secondary
sides are tuned to the same natural frequency, resonant coupling is achieved, maximising the power transfer capacity [73]. The efficiency of an IPT system can be boosted significantly using resonant compensation in the receiver, using a series or parallel connected capacitor similar to that in Figure 5.1 [73].

A series tuned pick-up delivers a constant voltage while a parallel tuned pick-up delivers a constant current. Figure 5.9 compares the output voltage and current characteristics of un-compensated, series tuned and parallel tuned pick-ups. Tuning the track and pick-up to the same frequency also helps filter out the undesired harmonics [73].

![Figure 5.9: Output characteristics of pick-up design [73]](image)

When the pick-up coil is tuned to the track’s operating frequency, maximum power transfer occurs and this power $P_m$ can be expressed as in (5.14) where $V_{oc}$ is the open circuit voltage and $I_{sc}$ is the short circuit current on the secondary side [73], [102]:

$$P_m = \frac{\omega I_p^2 M^2 Q_s}{L_s}$$  \hspace{1cm} (5.14)

Table 5.2 outlines the output properties of each pick-up compensation topology.

<table>
<thead>
<tr>
<th>Maximum outputs</th>
<th>Uncompensated</th>
<th>Series tuned</th>
<th>Parallel tuned</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Voltage $V_R$</strong></td>
<td>$V_{oc}/\sqrt{2}$</td>
<td>$V_{oc}$</td>
<td>$Q_s V_{oc}$</td>
</tr>
<tr>
<td><strong>Current $I_R$</strong></td>
<td>$I_{sc}/\sqrt{2}$</td>
<td>$Q_s I_{sc}$</td>
<td>$I_{sc}$</td>
</tr>
<tr>
<td><strong>Power $P_m$</strong></td>
<td>$V_{oc} I_{sc}/2$</td>
<td>$Q_s V_{oc} I_{sc}$</td>
<td>$Q_s V_{oc} I_{sc}$</td>
</tr>
</tbody>
</table>
Tuning both sides of the resonant circuit to the same resonance frequency $f_0$, also results in the currents being approximately sinusoidal [100]. The maximum efficiency of the power transfer is ultimately dependent upon the magnetic coupling between the coils, $k$, and the inductor quality factor $Q$ which increases with the operating frequency and the inductance and decreases with the resistance of the inductor [100]. As the secondary output needed to maintain a constant voltage for stable energy buffering, a series tuned topology was selected for the microactuator.

### 5.3.2.3. Quality Factor

In a resonant circuit, the quality factor, $Q$, is a measure of the “goodness” of the resonant circuit. A higher $Q$ indicates a narrower bandwidth with respect to the centre frequency, which is highly desired in many applications. The level of underdamping in the circuit is quantified by $Q$ where a higher $Q$ corresponds to a lower rate of energy loss relative to the stored energy of the resonator meaning the oscillations die out more slowly.

The quality factor of each side can be determined by (5. 1) and (5. 2) depending on the compensation topology. This can be further expresses as (5. 15) and (5. 16) for series and parallel tuning respectively, where $\omega_0$ is the angular resonant frequency.

\[
Q = \frac{R}{\omega_0 L} \quad (5. 15) \quad Q = \frac{\omega_0 L}{R} \quad (5. 16)
\]

The combined $Q$ for the IPT system is given by (5. 17), where $Q_p$ and $Q_s$ are the primary and secondary quality factors.

\[
Q = \sqrt{Q_p \cdot Q_s} \quad (5. 17)
\]

With a parallel tuned track and a series tuned pick-up in the ABL platform, $Q$ can be expressed by (5. 18).

\[
Q = \frac{R_s}{R_p} \sqrt{\frac{L_p C_s}{L_s C_p}} \quad (5. 18)
\]
The quality factor $Q$ and coupling factor $k$ determine the maximum achievable efficiency $\eta_{\text{max}}$ of the IPT system as shown in (5. 19) [100], which can be simplified further for high value of $k$ and $Q$ as in (5. 20). However if $Q$ is too high, due to the narrower bandwidth tuning of the pick up becomes challenging, therefore in IPT designs $Q \leq 10$ is maintained.

$$\eta_{\text{max}} = \frac{(kQ)^2}{(1 + \sqrt{1 + (kQ)^2})^2} \quad (5.19)$$

$$\eta_{\text{max}} \approx 1 - \frac{2}{kQ} \quad (5.20)$$

### 5.3.4. Primary Converter Design

The DC power input needs to be converted to AC, prior to entering the resonant tank on the primary side, which is done by the primary converter. Two types of primary converters were investigated in the ABL platform, to identify the more optimal solution.

#### 5.3.3.1. Primary Converter Topologies

Primary resonant converters can either be linear amplifiers or switch-mode converters [73]. Linear amplifiers are preferred when the power level is lower and a high quality sinusoidal output is vital, as the semiconductors operate in the linear region. Switch-mode converters are more commonly used in IPT applications where the power levels are in the kW order. The switching converter can be voltage or current fed [73].

Figure 5. 10 classifies the DC-AC power converter topologies.
Each of these converters have different features, pros and cons and limitations, which are summarised in Table 5.3.

**Table 5.3: Power converter topology comparison**

<table>
<thead>
<tr>
<th>Topology</th>
<th>Advantages</th>
<th>Disadvantages</th>
</tr>
</thead>
<tbody>
<tr>
<td>Voltage-fed full-bridge converter</td>
<td>• Track current can be regulated by duty cycle control</td>
<td>• 4 switching devices → more losses</td>
</tr>
<tr>
<td></td>
<td>• Compact and light due to lack of transformer</td>
<td>• Additional safety measures required</td>
</tr>
<tr>
<td></td>
<td>• More efficient as no transformer losses</td>
<td>• Not compatible with modules that must be earthed</td>
</tr>
<tr>
<td>Voltage-fed half-bridge converter</td>
<td>• 2 switching devices → lower switching losses</td>
<td>• Requires positive and negative voltage with respect to output neutral</td>
</tr>
<tr>
<td></td>
<td>• Control drives have common ground</td>
<td>• Lower output voltage</td>
</tr>
<tr>
<td>Current-fed full-bridge converter</td>
<td>• Inexpensive compared to current-fed push pull</td>
<td>• 4 switching devices → more losses</td>
</tr>
<tr>
<td></td>
<td></td>
<td>• Complex gate drives with isolation requirements</td>
</tr>
<tr>
<td>Current-fed push-pull converter</td>
<td>• 2 switching devices → lower switching losses</td>
<td>• Phase-splitting transformer more bulky and expensive</td>
</tr>
<tr>
<td></td>
<td>• Does not require isolated high-side gate drives</td>
<td>• Strong magnetic coupling required in split transformer to reduce leakage inductance</td>
</tr>
<tr>
<td></td>
<td>• Doubles resonant voltage</td>
<td></td>
</tr>
<tr>
<td></td>
<td>• Only one switch conducts at a time → less switching losses</td>
<td></td>
</tr>
<tr>
<td></td>
<td>• Control drives have common ground</td>
<td></td>
</tr>
</tbody>
</table>
### 5.3.3.2. Analysis of Current-fed Autonomous Push-Pull Converter

A current-fed autonomous push pull converter [103], as shown in Figure 5.11, was designed to deliver power to the microactuator. The DC supply voltage is connected to an inductor, which characteristically keeps the input current $I_{DC}$ into the converter constant. The input inductors $L_1$ and $L_2$ divides this DC current evenly to the two sides to form an approximate square wave input. The resonant tank consists of the track inductance $L_p$ and parallel connected capacitor $C_p$. The switches $S_1$ and $S_2$ are turned on alternatively to provide a closed path for the current through the primary track inductor $L_p$ in alternate directions. The series connected variable resistor $R_p$, represents the load of the system.

---

<table>
<thead>
<tr>
<th>Class D inverter</th>
<th>Class DE inverter</th>
<th>Class E inverter</th>
</tr>
</thead>
<tbody>
<tr>
<td>• Ideal for compact high-power applications</td>
<td>• Ideal for compact high-power applications</td>
<td>• Operable at higher frequencies than class D</td>
</tr>
<tr>
<td>• Reduced power waste as heat</td>
<td>• Reduced power waste as heat</td>
<td>• Complex circuitry</td>
</tr>
<tr>
<td>• Lower cost, size and weight of amplifier due to smaller (or no) heat sinks, and compact circuitry</td>
<td>• Lower cost, size and weight of amplifier due to smaller (or no) heat sinks, and compact circuitry</td>
<td>• Complex design</td>
</tr>
<tr>
<td>• Very high power conversion efficiency → theoretically 100%</td>
<td>• Very high power conversion efficiency → theoretically 100%</td>
<td>• High switching losses</td>
</tr>
<tr>
<td></td>
<td>• Complex circuitry</td>
<td>• Complex design</td>
</tr>
<tr>
<td>• Peak switch voltage is limited to the DC input voltage</td>
<td>• Reduced power waste as heat</td>
<td>• Operates more efficiently than either Class D or E</td>
</tr>
<tr>
<td>• Bandwidth theoretically bound by harmonic filter</td>
<td>• Bandwidth theoretically bound by harmonic filter</td>
<td>• Strict switching conditions</td>
</tr>
<tr>
<td></td>
<td></td>
<td>• Complex design</td>
</tr>
</tbody>
</table>

Due to maturity and common usage in IPT systems a switch-mode approach was chosen over the more complex linear amplifiers options. An autonomous current-fed push pull converter was initially designed for the ABL platform. A second voltage-fed full bridge converter was also designed to overcome the frequency instability issues associated with autonomous converter.
The drain source current of each switch $i_{DS1}$ and $i_{DS2}$ is given by (5.21) and (5.22).

$$i_{DS1}(t) = \begin{cases} I_{DC}, & 0 \leq \omega t < \pi \\ 0, & \pi \leq \omega t < 2\pi \end{cases} \quad (5.21)$$

$$i_{DS2}(t) = \begin{cases} 0, & 0 \leq \omega t < \pi \\ I_{DC}, & \pi \leq \omega t < 2\pi \end{cases} \quad (5.22)$$

The current through the resonant tank $I_T$ is given by (5.23).

$$I_T = i_{DS1}(t) - i_{DS2}(t) = \begin{cases} I_{DC}, & 0 \leq \omega t < \pi \\ -I_{DC}, & \pi \leq \omega t < 2\pi \end{cases} \quad (5.23)$$

Each gate is driven by feeding the resonant voltage of the opposite side, with each side identical and is outlined in Figure 5.12. Traditional gate driver ICs have not been used in the autonomous converter. The gate drive operation can be analysed as follows.
The autonomous gate drive signal is provided by feeding back the resonant tank voltage to the gate of $S_1$ and $S_2$ via cross connected diodes $D_2$ and $D_1$ respectively, in Figure 5.12. The gate voltage $V_{G1}$ is provided by the DC source voltage $V_{DC}$ and the resonant tank voltage at point B, $V_B$. $V_{DC}$ is connected to the gates through the current limiting resistors $R_1$ and $R_2$. When the resonant voltage $V_P$ is high at A, the capacitor $C_{2s}$ is charged to turn $S_2$ on. This drives the voltage at point B to zero which causes $D_1$ to become forward biased and turn off $S_1$. As the voltage through the resonant tank starts to decline at point A during the second half cycle, diode $D_2$ becomes forward biased ideally making the gate voltage of $S_2$ and hence turn it off. The simultaneously increasing voltage at B is fed back to $S_1$ causing it to turn it on. The low state of the gate drive signal does not go exactly to zero due to the forward bias voltage drop across the diodes, which ranges between 0.2-0.7 V.

The operation of the gate drive circuit can be further analysed mathematically using the superposition theorem. The first instance when $V_{DC}$ is the only source, the circuit is simplified as seen in Figure 5.13 (a). Figure 5.13 (b) shows the second instance where $V_B$ is the only source.

\[ V_{G1}(t) = V_{DC} \left( 1 - e^{-t/ R_1(C_{iss} + C_s)} \right) \]  \hspace{1cm} (5.24)

Figure 5.13: Gate driver circuit with superposition theorem applied

From Figure 5.13 (a) the gate voltage can be derived as in (5.24).

\[ I_s = I_R + I_{G1} \Rightarrow C_s \frac{dV_{CS1}}{dt} = C_{iss} \frac{dV_{G1}}{dt} + \frac{V_{G1}}{R_1} \]  \hspace{1cm} (5.25)
\[
\frac{dV_{G1}}{dt} + \frac{1}{R_1(C_S + C_{iSS})} V_{G1} = \frac{C_S}{(C_S + C_{iSS})} \frac{dV_g}{dt} \quad (5.26)
\]

The half wave voltage across the resonant tank \( V_B \) is given by (5.27).

\[
V_B(t) = \begin{cases} 
\pi V_{DC} \sin \omega t & 0 \leq t < \pi \\
0 & \pi \leq t < 2\pi 
\end{cases} 
\quad (5.27)
\]

The governing equation for the gate voltage can then be written as (5.28).

\[
\frac{dV_{G1}}{dt} + \frac{1}{R_1(C_S + C_{iSS})} V_{G1} = \frac{C_S}{(C_S + C_{iSS})} \omega \pi V_{DC} \cos \omega t 
\quad (5.28)
\]

The natural and forced response for the gate voltage is combined to express the total solution for \( V_{G1} \) as in (5.29).

\[
V_{G1} = \left\{ Ke^{-t/\tau}\right\} + \left\{ A \cos \omega t + B \sin \omega t \right\} 
\quad (5.29)
\]

The time constant \( \tau \) is expressed as in (5.30), while the constant \( K \) of the natural response can be found by applying initial conditions, once the forced response has been determined.

\[
\tau = R_1(C_S + C_{iSS}) 
\quad (5.30)
\]

Combining equations (5.28) and the forced response gives (5.31) and (5.32).

\[
A = \frac{\pi V_{DC} C_{S1} R_1 \omega}{(1 + \tau^2 \omega^2)} 
\quad (5.31)
\]
\[
B = \tau \omega A = \frac{\pi V_{DC} C_{S1} R_1 \tau \omega^2}{(1 + \tau^2 \omega^2)} 
\quad (5.32)
\]

Using the values for constants \( A \) and \( B \) together with the initial condition \( V_{G1} = 0 \) and \( t = 0 \) give the value for \( K \) according to (5.33).

\[
V_{G1}(0) = Ke^{0/\tau} + A \cos 0 + B \sin 0 \quad \Rightarrow \quad K = -A = -\frac{\pi V_{DC} C_{S1} R_1 \omega}{(1 + \tau^2 \omega^2)} 
\quad (5.33)
\]

The complete solution governing for the gate drive voltage is then given by (5.34).
\[ V_{G1}(t) = V_{DC} \left( 1 - e^{-t/\tau} \right) + \left( V_{DC} \frac{\pi C_{S1} R_1 \omega}{(1 + \frac{\tau^2}{\omega^2})} \right) \left[ (\cos \omega t + \tau \omega \sin \omega t) - e^{-t/\tau} \right] \] (5.34)

The gate limiting resistors and speed up capacitors should be designed to achieve the desired turn on speed. The threshold \( V_{GS} \) must not be exceeded for the switch, and the Zener diodes with \( V_z < V_{GS} \) ensures this. The current limiting resistors also prevent the shorting of the DC source. The resistor capacitor combination forms an RC circuit which mainly determines the turn on speed of the switches. For a switch with a given input capacitance \( C_{iss} \), a smaller resistance will give a smaller time constant to charge the capacitor faster. But due to the larger current drawn, the losses will increase. Adding the speed up capacitors \( C_{S1} \) and \( C_{S2} \), some charging current will be drawn from the resonant voltage to speed up the turn on process. This allows for a higher \( R_1 \) and \( R_2 \) to minimise switching losses.

The steady state analysis and zero voltage switching (ZVS) are the primary factors that govern the smooth operation of the autonomous converter. The polarity of the voltage across the resonant capacitor governs the switching operation. To ensure ZVS one switch is turned on while the other switch is off. The current flow through the resonant tank is shown for the two different half cycles, in Figure 5.14. To avoid complexity the gate drive details have been eliminated in this circuit.

![Figure 5.14: Operating cycle of the autonomous push pull converter](image-url)
The steady state voltage across the capacitor can be expressed by (5.35), where $V_{\text{rms}}$ is the rms voltage across the capacitor and $\omega$ is the zero voltage switching frequency.

$$v_c(t) = \sqrt{2}V_{\text{c(rms)}} \sin \omega t \quad (5.35)$$

The voltage formed across each corresponding switch is a positive half sinusoid of the capacitor voltage. The peak voltage $V_{A(\text{peak})}$ (or $V_{B(\text{peak})}$) formed across a single switch for a full period $T$, when the average DC voltage delivered by the supply is $V_{\text{DC}}$, is given by (5.36):

$$V_{\text{DC}} = \frac{1}{T} \int_0^T v_c(t) \, dt = \frac{1}{T} \int_0^T \sqrt{2}V_{\text{c(rms)}} \sin \omega t \, dt \quad \Rightarrow \quad V_{\text{c(rms)}} = \frac{\pi V_{\text{DC}}}{\sqrt{2}}$$

$$|V_{A(\text{peak})}| = |V_{B(\text{peak})}| = |V_{\text{c(peak)}}| = \pi V_{\text{DC}} \quad (5.36)$$

The current $I_p$ through the primary inductor $L_p$ can be determined as shown in (5.37), which is further simplified using $Q$, given by (5.16) since the practical operating frequency is taken as the un-damped natural frequency defined by (5.13):

$$I_p = \frac{\pi V_{\text{DC}}}{\sqrt{(\omega L)^2 + R_p^2}} \approx \frac{\pi V_{\text{DC}}}{(\omega_0 L)\sqrt{Q^2 + 1}} \quad (5.37)$$

The alternating operation of the switches generate a continuous sinusoidal voltage/current $V_{AB}$ through the primary track inductor as illustrated in Figure 5.15.
Previously it was approximated that the operating frequency, or ZVS frequency $\omega$ is approximately same as the undamped natural frequency $\omega_0$. The accurate ZVS frequency is determined by the analysing one half cycle of the converter operation under steady state conditions, as shown in Figure 5. 16. In one half cycle inductor $L_1$ becomes parallel with the source, while $L_2$ is in series with the resonant tank. The input current $I_{DC}$ is evenly split by these inductors as they are much larger in comparison with the resonant inductor $L_P$.

![Figure 5. 16: Steady state analysis for ZVS switching frequency](image)

Figure 5. 15: Resonant tank and gate voltages
The resonant voltage $V_c$ can be determined using equations (5.38) derived from the voltage formed across the resonant capacitor.

$$V_c = \frac{1}{C_p} \int I_c \, dt = \frac{1}{C_p} \int \left(I_p + \frac{I_{DC}}{2}\right) \, dt$$

$$L_p C_p \frac{d^2 V_c}{dt^2} + R_p C_p \frac{d V_c}{dt} + V_c = \frac{I_{DC}}{2} R_p$$  \hspace{1cm} (5.38)

Using the initial conditions for ZVS gives the complete solution as in (5.39) where the initial phase angle of the voltage $\theta_v$ and time constant $\tau$ are defined by (5.40) and (5.41).

$$V_c(0) = 0 \quad \text{and} \quad \frac{d V_c(0)}{dt} = \frac{1}{C_p} \left(I_p(0) + \frac{I_{DC}}{2}\right)$$

$$V_c(t) = \frac{I_{DC} R_p}{2 \sin \theta_v} e^{-t/\tau} \sin\left(\omega_f t - \theta_v\right) + \frac{I_{DC}}{2} R_p$$  \hspace{1cm} (5.39)

$$\theta_v = \tan^{-1} \left[ \frac{\omega_f I_{DC} R_p C_p}{L_p P(0) + I_p \left(2 L_P - R_p^2 C_P\right)} \right]$$  \hspace{1cm} (5.40)

$$\tau = \frac{2L}{R}$$  \hspace{1cm} (5.41)

As the voltage goes to zero every half cycle, if the switching period is $T_s$, at $t = T_s/2$ equation (5.42) can be derived.

$$e^{-\left(T_s/2\right)} \sin\left(\frac{\omega_f T_s}{2} - \theta_v\right) + \sin \theta_v = 0$$  \hspace{1cm} (5.42)

The resonant current $I_p$ can be determined using equation (5.43) which is derived using the current through the inductor and voltage across capacitor. The voltages across inductor and resistor are $V_{LP}$ and $V_{RP}$ respectively.

$$V_c = \frac{1}{C_p} \int I_c \, dt = V_{LP} + V_{RP} = L_p \frac{dI_p}{dt} + R_p I_p$$

$$L_p C_p \frac{d^2 I_p}{dt^2} + R_p C_p \frac{d I_p}{dt} + I_p = -I_{DC}$$  \hspace{1cm} (5.43)

The complete solution for the resonant current is derived as shown in (5.44) where the initial phase angle of the current $\theta_i$ is given by (5.45).
\[ I_p(t) = \frac{2I_p(0) + I_{DC}}{2 \sin \theta_i} e^{-\frac{t}{\tau}} \sin(\omega_f t + \theta_i) - \frac{I_{DC}}{2} \]  
\[ \theta_i = \tan^{-1} \left[ \frac{2 \omega_f L_p (I_{DC} + 2I_p(0))}{R_p [I_{DC} - 2I_p(0)]} \right] \]  
\[ 5.44 \]  
\[ 5.45 \]

The resonant current changes polarity at the end of each half cycle, therefore \( I_p(T_s/2) = -I_p(0) \) giving (5.46).
\[ e^{-\frac{(T_s/2)}{2\tau}} \sin \left( \frac{\omega_f T_s}{2} + \theta_i \right) - \left( \frac{I_{DC}}{2I_p(0) + I_{DC}} \right) \sin \theta_i = 0 \]  
\[ 5.46 \]

Solving for \( T_s \) allows the zero voltage switching frequency to be determined, but due to the non-linearity of (5.44) and (5.45) these cannot be solved using analytical methods. But numerically solving for \( T_s \) allows the switching frequency to be determined.

Since the converter has autonomous driving, it is important to analyse the start up mechanism of the converter. In the autonomous design theoretically an initial energy is required to start up the converter, which would technically require energy injection into either the resonant capacitor, resonant inductor or the DC inductor. This requires extra charging circuitry leading to higher costs and complexity. But one outstanding practical feature is the autonomous regenerative process that starts the converter up, which can be explained as follows.

The two switches are initially off, and once the DC source is turned on both switches will tend to turn on by \( V_{DC} \) and gate resistors \( R_1 \) and \( R_2 \). But turning one switch on will short the gate of the other switch forcing it to turn off. Due to practical non-identicalities of the switches one switch will fully turn on and turn the other switch off, which will the start up the process of alternatively turning \( S_1 \) and \( S_2 \) on and off. Bi-stable circuit oscillation and zero voltage switching is established in the circuit.

The mathematical equations that govern the process during the start up time \( t_0 \) can be given by (5.47) - (5.49), where \( L_{in} \) is the equivalent DC inductance seen by the supply voltage.
\[
L_{\text{in}} \frac{dI_{\text{DC}}(t)}{dt} = V_{\text{DC}} \quad (5.47)
\]

\[
L_{\text{in}} = \frac{L_1}{L_2} = \frac{L_1}{2} \quad (5.48)
\]

\[
I_{\text{DC}}(t) = \frac{V_{\text{DC}}}{L_{\text{in}}} t_0 \quad (5.49)
\]

The operation of the converter was simulated on LT Spice with the following circuit parameters shown in Table 5.4.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>(L_P)</td>
<td>5.7 µH</td>
</tr>
<tr>
<td>(C_P)</td>
<td>330 pF</td>
</tr>
<tr>
<td>(L_1 = L_2)</td>
<td>1 mH</td>
</tr>
<tr>
<td>(V_{\text{DC}})</td>
<td>15 V</td>
</tr>
</tbody>
</table>

The resonant voltage, current and gate drive voltages obtained from the simulations are shown in Figure 5.17. The full resonant voltage and current are shown in the first plot while the following plots shown the resonant voltage and gate drive voltage for each side.
5.3.3.3. Voltage-fed Full Bridge Converter

Due to frequency instabilities associated with the autonomous push-pull converter a voltage fed full bridge converter was also designed, as shown in Figure 5. 18. In the configuration, the compensation capacitor $C_P$ cannot be parallel tuned as there is a possibility of shorting the voltage source inverter output $V_P$ when the $C_P$ is fully discharged. Therefore, to avoid any overcurrent damage a series tuned track is used for the voltage fed converters. The four switches are turned on in pairs as $S_1$ together with $S_4$ and $S_2$ together with $S_3$. 

Figure 5.17: Simulation results of current fed push-pull converter
The switches are operated at a predefined frequency of the resonant tank. If the switches are operated exactly at zero current crossing (ZCS) the output current and voltage are in phase, and generate zero switching losses, as shown in Figure 5.19 (a). Due to practical difficulties in achieving ZCS, practical circuit shows a leading or lagging angle between the square wave input voltage and output current, as seen in Figure 5.19 (b) and (c). Although implementing fixed frequency control is easier and straightforward, hard switching results in high switching losses and electromagnetic interference (EMI). Controlling the magnitude of the output current is also another difficulty associated with fixed frequency control and hence phase shift control, as shown in Figure 5.19 (d), is more desirable.

In order to achieve phase shift control the phase angle $\phi$ of either the upper or lower pair of switches is controlled. If the pair of switches $S_1$ and $S_3$ are completely out of phase with a phase shift of $180^\circ$ the input voltage is maximized. If the same switches are in phase with a phase angle of $0^\circ$ the input voltage applied to the resonant tank becomes zero driving the track current to zero eventually. By maintaining continuous phase shift control the magnitude of the output current can be regulated. The relationship between the input DC voltage $V_{DC}$ and the input current to the resonant tank $I_P$ can be expressed by (5.50).

$$V_P = \frac{4}{\pi \sqrt{2}} V_{DC} \sin \left(\frac{\phi}{2}\right) \quad (5.50)$$
As \( S_1 \) and \( S_4 \) are turned on after the zero current crossing and \( S_2 \) and \( S_3 \) are switched off before the same point turn-on turn off switching losses are zero. Charging and discharging parallel soft switching capacitors is used to achieve soft switching. The phase shifted PWM control of the switches is done using an FPGA. The NI 9401 CompactRio controller, coupled with a NI 9401 digital module and NI 9221 voltage input module, shown in Figure 5. 20, is used to control the H bridge driver. A SM72295 full bridge driver IC was controlled by the FPGA to drive the full bridge converter.
The primary coil $L_p$ impedance is designed to be slightly higher than that of the resonant capacitor $C_p$ to achieve the desired primary current $I_p$, which is given by (5.51).

$$I_p = \frac{V_p}{R_p + j\omega L_p + \frac{1}{j\omega C_p}}$$

(5.51)

The only major disadvantage is the full primary current circulates through each switch leading to high conduction losses and therefore lower efficiency. As the primary current is directly dependent on the total impedance, the current is highly sensitive to slight variations in the impedance.

The H-bridge converter was simulated in LT spice with the following circuit parameters shown in Table 5.5.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_p$</td>
<td>5.7 μH</td>
</tr>
<tr>
<td>$C_p$</td>
<td>27 nF</td>
</tr>
<tr>
<td>$V_{DC}$</td>
<td>15 V</td>
</tr>
<tr>
<td>$V_{GS}$</td>
<td>10 V (square wave)</td>
</tr>
</tbody>
</table>
The resonant voltage, current and gate drive voltages are shown in Figure 5. 21. The first plot shows the resonant tank voltage and current, while the following plots illustrate the resonant voltage on each side, along with the respective gate drive voltage.

![Figure 5. 21: LT spice simulation results for voltage fed full bridge converter](image)

5.3.5. Secondary Pick-Up Circuit Design

The secondary pick up circuit, shown in Figure 5. 22, is combined with energy storage and valve control modules and encapsulated within the microvalve actuator. The secondary pick up coil $L_s$ is coupled magnetically with the primary track and is series tuned with the compensation capacitor $C_s$ to the resonant frequency of the primary converter. The series tuned converter and pick up combination forms a constant current source which forms a constant AC current at the output of the pickup. This is ideal for
charging the supercapacitor energy buffer, where the AC output is rectified through the full bridge rectifier and Zener regulated for a constant voltage. The output of the supercapacitor buffer is dropped down using a resistor divider and regulated using Zener diodes. Stable output voltages of 15 V, 5 V and 3 V are generated for different parts of the actuator control circuit described in section 4.3.1.1.

The output from the supercapacitor buffer controls the current flow into the actuator circuit. This is controlled by the actuator control signal, where the grounds of both sides are made common when actuation is required, as outlined in Figure 5.22.

The secondary circuit was simulated in LT Spice, to show the pick up voltage, the rectified DC output, dropped down DC voltages and the energy buffer charge current in Figure 5.23.

![Diagram of the wireless power supply with supercapacitor energy buffer](image)
5.4. Energy Buffer Design

The momentary power consumption only during actuation and zero consumption between actuations is a key feature of the actuation technique described in Chapter 4. Although this mechanism is extremely energy efficient, the momentary power consumption of the actuator is around 40 W. Although a high power IPT system can be designed to accommodate this 40 W requirement, for most part of the operation this power is unutilised and will lead to unnecessary heating and require larger components to accommodate the high current levels. Therefore, the IPT power supply in this microactuator is accompanied by an energy buffer which stores the energy transferred over the IPT link. And the buffered energy is used to source the peak power requirements during actuation. The energy buffer is required to source a peak current of 2.5 A during actuation. The energy buffer also makes the actuator self-sustainable, and operate even when the IPT supply is offline.
5.4.1. Overview of Energy Storage Technologies

Several energy storage technologies such as rechargeable batteries and supercapacitors were investigated for feasibility in the microactuator energy buffer design. Figure 5.24 outlines some of the popular, rechargeable energy storage technologies.

Since its invention by Alexander Volta, batteries have evolved over the past 200 years with different chemical compositions and properties [104]. Rechargeable batteries or secondary batteries are more versatile and easy to adapt due to commercially available charging solutions. They can be recharged up to 1000 times depending on the battery type, very deep discharge shortens the cycle life, charge time varies between 1-12 hours which is dependent upon battery condition, depth of discharge and other factors. The drawbacks of utilizing rechargeable batteries in the microactuator include the limited lifetime, lower power capability, low energy-efficiency, and the need for additional specialised charging circuitry.

Figure 5.24: Rechargeable energy storage technologies
Fuel cells use externally stored chemical energy to generate electricity, but unlike with batteries the energy storage and power generation are separate. They require large control equipment such as fuel pumps, fuel tanks, cooling system etc. to operate, which makes them unsuitable for portable applications like the microactuator.

Capacitors store energy on the surface of the electrodes, separated by an electrolyte to generate an electric field, they are much smaller in volume and weight in comparison with fuel cells and most battery types. Electrolytic capacitors are a sub category of capacitors and a very mature, abundantly used technology. The micro-meter thickness metal oxide based electrodes are not very efficient, but capable of handling very high voltages, delivering high current for extremely short periods. A relatively new class of capacitors, known as supercapacitors, have much higher capacitances due to the activate carbon electrodes. Therefore, they are capable of storing larger amounts of energy, without compromising the physical dimensions. The lower breakdown voltage of solvents inside supercapacitors have resulted in lower operating voltages between 0.5-5.5 V. Supercapacitors have virtually unlimited cycle life and can have 90% or more efficiency. Their key feature is the higher power densities than batteries and hence the ability to source peak power requirements. However, they offer lower energy densities than batteries and fuel cells. In general capacitors do not require special charging circuitry, which significantly reduces the circuit complexity, cost and real estate.

Table 5.6 compares the key features of the discussed energy storage technologies.

| Property                        | Capacitors     | Supercapacitors | Batteries       | Fuel Cells     |
|---------------------------------|----------------|-----------------|-----------------|----------------|-----------------|-----------------|-----------------|-----------------|-----------------|-----------------|-----------------|-----------------|-----------------|-----------------|-----------------|-----------------|-----------------|-----------------|-----------------|-----------------|-----------------|
| Charge/Discharge Time           | ps - ms        | ms - s          | 1 - 10h         | 10 - 300h      |
| Operating Temperature           | -20 to +100°C  | -40 to +85°C    | -20 to +65°C    | +25 to +90°C   |
| Operating Voltage               | 6 - 800V       | 2.3 - 2.75V     | 1.25 - 4.2V     | 0.6V           |
| Capacitance                     | 10pF - 2.2mF   | 100mF - 1500F   | N/A             | N/A            |
| Life                            | >100,000 cycles| 50,000+ h, Unlimited cycles | 150 to 1,500 cycles | 1,500 to 10,000 h |
The Ragone plot in Figure 5.25 compares the power density against the energy density for these energy storage technologies. The energy density is defined as the amount of energy per unit weight Watt-hours per kilogram (Wh/kg) while the power density refers to the rate of energy release per unit weight measured in Watts per kilogram (W/kg) [106]. It is evident that capacitors have the highest power density and hence are the most suitable option to source the momentary peak current requirement of the microactuator, although batteries and fuel cells with their much higher energy densities could power the actuator for longer.

![Ragone Plot](image.png)

*Figure 5.25: Specific power against specific energy for rechargeable energy storage options [105]*

When considering the charge/discharge capabilities batteries offer slow and steady charge discharge profiles, while capacitors are charged and discharged much faster, as
shown in Figure 5. 26 [107]. As opposed to hours taken by batteries, supercapacitors charge in a matter of minutes.

Figure 5. 26: Charge-Discharge curve of battery and supercapacitor [108]

With the energy storage capability of batteries and rapid charge/discharge characteristics of capacitors, supercapacitors bridge the gap between conventional capacitors and rechargeable batteries [109]. Supercapacitors are better suited to meet the peak power requirement of the actuator, as they provide better power density than batteries and better energy density than capacitors. Disadvantages associated with batteries such as self-discharge, the ‘memory effect’ needing to fully discharge before recharging, toxicity and environmental impacts, requirement for sophisticated charging circuits etc. [108] are not experienced with supercapacitors.

5.4.2. Supercapacitor Technology

The main difference between conventional capacitors and supercapacitors is in their construction. The higher capacitance and energy storage capability in supercapacitors is achieved by the increased capacitor plate area using carbon nanotubes with extremely small separation distance [110].

Supercapacitors use different percentages of electrostatic double layer capacitance and electro-chemical pseudocapacitance, to makeup the total capacitance. Based on their electrode design supercapacitors are divided as electric double-layer capacitors (ELDC), pseudocapacitors and hybrid capacitors. ELDCs that use carbon electrodes or its...
derivatives have higher static double-layer capacitance. The charge separation is between 0.3-0.8 nm [108]. These are the least costly and most commonly available. Electrochemical pseudocapacitors use metal oxide or conducting polymer based electrodes which result in higher electrochemical pseudocapacitances over the double-layer capacitance. Hybrid capacitors have asymmetric electrodes, one mostly electrostatic and the other mostly electrochemical capacitance. The lithium-ion capacitor is common example of the hybrid capacitor.

Figure 5. 27 shows the internal arrangement of the layers within the supercapacitor.

The capacitance of supercapacitors varies between a few Farads to a few thousand Farads, while conventional capacitors have capacitances in the picofarad (pF) – microfarad (μF) range [107]. Of all capacitor types, supercapacitors have the highest available capacitance values per unit volume and the greatest energy density [112]. Although supercapacitors were previously expensive, research and development has produced cheaper, low profile devices with low equivalent series resistance (ESR), chemical and physical stability [113]. One major constraint of available supercapacitors is the low voltage rating varying between 2.3 – 5.5 V [107]. However, to achieve higher voltages, series combinations of supercapacitors can be used. Figure 5. 28 shows some commercially available supercapacitor.
5.4.3. Supercapacitor Energy Buffer

5.3.3.4. Theoretical Modelling of Supercapacitor Energy Buffer

As energy storage and peak power delivery are the main tasks of the supercapacitor the energy and power capabilities of the supercapacitor energy buffer is modelled.

In any capacitor, the capacitance $C$ is a function of the electrode surface area $A$, separation distance $d$ and the permittivity of dielectric medium $\varepsilon$ as given by (5.52).

$$C = \frac{\varepsilon A}{d}$$

(5.52)

The supercapacitor energy buffer, as with any other energy storage module, can be modelled using its internal resistance and its ideal voltage source, and is connected to the actuator circuit, as shown in Figure 5.29.

![Electric circuit model of supercapacitor](image)
The voltage at the actuator $V_{\text{act}}$ is given by (5.53), where $R_{\text{act}}$ is the resistance seen at the actuator, $R_{\text{ESR}}$ is the equivalent series resistance (ESR) or the internal resistance of the supercapacitor and $V_{\text{super}}$ is the ideal source voltage given by the supercapacitor [117].

$$V_{\text{act}} = \frac{V_{\text{super}} R_{\text{act}}}{R_{\text{act}} + R_{\text{ESR}}} \quad (5.53)$$

The output current $I_{\text{act}}$ of the energy buffer is given by (5.54):

$$I_{\text{act}} = \frac{V_{\text{super}}}{R_{\text{act}} + R_{\text{ESR}}} \quad (5.54)$$

The power delivered from energy buffer to the actuator $P_{\text{act}}$ is given by (5.55):

$$P_{\text{act}} = V_{\text{act}} I_{\text{act}} = \left(\frac{V_{\text{super}} R_{\text{act}}}{R_{\text{act}} + R_{\text{ESR}}}\right) \left(\frac{V_{\text{super}}}{R_{\text{act}} + R_{\text{ESR}}}\right)$$

$$P_{\text{act}} = V_{\text{act}} I_{\text{act}} = \frac{V_{\text{super}}^2}{\left(\frac{R_{\text{ESR}}}{R_{\text{act}}} + 1\right)^2}$$

$$P_{\text{act}} = V_{\text{act}} I_{\text{act}} = \frac{V_{\text{super}}^2}{4 R_{\text{ESR}}} \quad (5.55)$$

The maximum power $P_{\text{max}}$ is transferred if $R_{\text{act}} = R_{\text{ESR}}$, according the maximum power transfer theorem [118]. This is given by (5.56), showing that the lower the ESR the higher the power transfer capability.

$$P_{\text{max}} = \frac{V_{\text{super}}^2}{4 R_{\text{ESR}}} \quad (5.56)$$

The power density of a supercapacitor is defined as the maximum power transferable per unit weight or unit volume. Therefore, the power density $P_{\text{density}}$ is given by (5.57) where the $w$ is the weight of the supercapacitor.

$$P_{\text{density}} = \frac{V_{\text{super}}^2}{4 R_{\text{ESR}} w} \quad (5.57)$$

This shows how the very low ESR of supercapacitors, ranging between 0.3 mΩ to 10s of milli-Ohms [117], allows for this much larger power density as shown in Figure 5.25. The power density of supercapacitors is only lower than conventional capacitors because it is limited by a much lower voltage rating.
The internal resistance of a supercapacitor against the depth of discharge is compared against that of a battery in Figure 5.30.

![Figure 5.30: Internal resistance against depth of discharge of a supercapacitor and battery [117]](image)

This significant increase in internal resistance as a battery discharges, is what makes it unsuitable to meet peak power requirements as there is an exponential drop in its power density. With their consistent and very low internal resistance, the power density of supercapacitors remains high and constant as it discharges making it much more suitable for peak power needs.

The actuator requires approximately 2.5 mJ per actuation and zero energy between actuations. The energy storage capability $E$ of a supercapacitor is dependent on the voltage $V$ and capacitance $C$, and is defined by (5.58).

$$ E = \frac{1}{2} CV^2 \quad (5.58) $$

The energy density $E_{density}$ is defined as energy stored per unit weight or volume, and is given by (5.59). It shows that higher voltage ratings and capacitances lead to higher
energy densities. Conventional capacitors have higher voltage ratings, but much lower capacitances. The much higher capacitances in supercapacitors compensate for their lower voltage ratings, as they evidently have higher energy densities than capacitors as seen in Figure 5.25.

\[ E_{\text{density}} = \frac{CV_{\text{super}}^2}{2w} \quad (5.59) \]

### 5.3.3.5. Electrical Circuit Design of Energy Buffer

To source the large coil current for actuation, approximately 15 V were required by the actuator. To source 2.5 mJ energy at the required voltage of 15 V the capacitance required is approximately 20 µF according to (5.58). As supercapacitors provide much higher capacitances, a series combination of three 2 F/5.5 V capacitors, as shown in Figure 5.31, an energy reserve of 240 J. Therefore, thousands of actuations can be sourced before the supercapacitor needs to be fully charged. The series parallel combination is used to make up for the capacitance reduction due to connecting the capacitors in parallel. Zener regulation is used to maintain a constant output voltage of 15 V.

![Figure 5.31: Series connected supercapacitor energy buffer](image)

The lower output voltages sometimes produced by the IPT secondary pickup results is a lower DC voltage to charge supercapacitor bank. This results in the capacitor bank not being fully charged to 15 V and hence be unable to source the actuation current. A second energy buffer design as shown in Figure 5.32 was investigated, where the charging and discharging of the supercapacitors were done in different configurations.
During the charging cycle $S_1$, $S_3$, $S_4$ and $S_6$ are closed while $S_2$ and $S_5$ are opened making the capacitors parallel. The maximum DC charging voltage required is 5.5 V, which is the rated voltage of a single capacitor. Therefore, the IPT output can be lower. During the discharging cycle $S_2$ and $S_5$ are closed while $S_1$, $S_3$, $S_4$ and $S_6$ are opened making the capacitors series to generate a higher voltage at the output of the energy buffer. This arrangement allows for a lower IPT input yet a higher energy buffer output voltage to source the actuations. Control mechanisms configure the same fully charged supercapacitors to a series arrangement. But due to complexity and difficulties in driving the high side switches, it was decided to retain the design shown in Figure 5. 31 with a higher IPT output.

### 5.3.3.6. Simulation Study of Supercapacitor Energy Buffer

The energy buffer charging cycle was simulated using LT spice as shown in Figure 5. 33.
5.5. Wireless Power Platform Development

An initial prototype of the wireless power supply was created using through hole components and later redesigned with a more energy efficient SMD version. The track was implemented with Litz wire, a multi strand wire designed to minimise the skin effect of the current. The pickup coil is a commercially available with a 7.2 µH, 6 mm radius charging coil by Wurth Elektronik, as illustrated in Figure 5.34.

![Figure 5.34: Track and coil setup of wireless power supply](image)

The initial implementation of the autonomous push-pull primary converter is shown in Figure 5.35. It uses IRF510 MOSFETs for driving the current.
The H-bridge primary converter, which was more suited for the communication needs of the system is shown in Figure 5.36. It is designed using a TI SM72295 H-bridge driver and four IRF510 MOSFET switches.

The secondary pick-up circuit is illustrated in Figure 5.37, which uses the secondary coil, compensated by 27 nF capacitor to achieve resonance at the 800 kHz operating frequency.
The supercapacitor energy buffer was constructed using 5.5 V rated, 0.22 F supercapacitors in series as shown in Figure 5. 38. The total equivalent capacitance with 3 series capacitors is 0.073 F, with an energy storage capability of 8.79 J.

These proof of concept prototypes can further be optimised for space to be enclosed within the microvalve, which is one of the future works suggested in Chapter 8.
5.6. Experimental Results

The wireless power supply was tested using both full-bridge and autonomous push-pull converters and different pick up coils with the following specifications shown in Table 5.7.

Table 5.7: Performance summary of different pick up coils

<table>
<thead>
<tr>
<th></th>
<th>Coil 1</th>
<th>Coil 2</th>
<th>Coil 3</th>
</tr>
</thead>
<tbody>
<tr>
<td>Manufacturer</td>
<td>Würth Elektronik</td>
<td>Würth Elektronik</td>
<td>Würth Elektronik</td>
</tr>
<tr>
<td>Inductance (marked)</td>
<td>7.2 µH</td>
<td>24.2 µH</td>
<td>16.7 µH</td>
</tr>
<tr>
<td>Inductance (measured)</td>
<td>7.4 µH</td>
<td>24.6 µH</td>
<td>15.5 µH</td>
</tr>
<tr>
<td>Shape</td>
<td>Circular</td>
<td>Circular</td>
<td>Rounded Rectangular</td>
</tr>
<tr>
<td>Coil dimensions</td>
<td>Ø 6 mm</td>
<td>Ø 7.2 mm</td>
<td>38.5 x 30.5 mm</td>
</tr>
<tr>
<td>ESR</td>
<td>0.4 Ω</td>
<td>1.2 Ω</td>
<td>0.25 Ω</td>
</tr>
<tr>
<td>Rated current</td>
<td>0.5 A</td>
<td>0.4 A</td>
<td>2 A</td>
</tr>
<tr>
<td>Saturation current</td>
<td>1 A</td>
<td>1 A</td>
<td>4 A</td>
</tr>
<tr>
<td>Q</td>
<td>10</td>
<td>13</td>
<td>32</td>
</tr>
</tbody>
</table>

5.6.1. Push Pull Converter Performance

The current fed autonomous push pull converter was also tested using different pick up combinations. The push pull converter stably operated around 3.5 MHz, but as there is very little control over the operating frequency on the primary side, significant variations in the frequency were observed, as shown in Figure 5.39. This made it difficult to tune the secondary pick up to obtain higher power outputs. In Figure 5.39 (b) it can be seen that the RMS output voltage drops significantly to 2.44 V from 4.16 V, when the pickup does not resonate with the converter.
The converter was capable of delivering up to 1.127 W to the actuator and the measured pick up power and output voltage are plotted against the primary track current in Figure 5.40.
5.6.2. Full Bridge Converter Performance

The implemented full bridge inverter was tested with the chosen secondary pick up coils at several frequencies. As the converter is externally driven, the frequency instabilities of the push pull converter were not experienced. At each frequency, the track needed to be tuned to this frequency to achieve resonant operation. The inductance and capacitance used to achieve resonant operation are shown in Table 5.8.

Table 5.8: Pick-up inductor capacitor combinations experimented

<table>
<thead>
<tr>
<th>Track Inductance</th>
<th>Track Capacitance</th>
<th>Pick Up Inductance</th>
<th>Pick Up Capacitance</th>
<th>Operating Frequency</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.5 µH</td>
<td>1.8 µF</td>
<td>7.2 µH</td>
<td>330 nF</td>
<td>100 kHz</td>
</tr>
<tr>
<td>1.5 µH</td>
<td>1.8 µF</td>
<td>24.2 µH</td>
<td>100 nF</td>
<td>100 kHz</td>
</tr>
<tr>
<td>1.5 µH</td>
<td>1.8 µF</td>
<td>15.5 µH</td>
<td>150 nF</td>
<td>100 kHz</td>
</tr>
<tr>
<td>1.5 µH</td>
<td>1.8 µF</td>
<td>5.7 µH</td>
<td>470 nF</td>
<td>100 kHz</td>
</tr>
<tr>
<td>5.7 µH</td>
<td>6.8 nF</td>
<td>7.2 µH</td>
<td>5.5 nF</td>
<td>800 kHz</td>
</tr>
<tr>
<td>5.7 µH</td>
<td>6.8 nF</td>
<td>24.2 µH</td>
<td>1.5 nF</td>
<td>800 kHz</td>
</tr>
<tr>
<td>5.7 µH</td>
<td>6.8 nF</td>
<td>15.5 µH</td>
<td>2.6 nF</td>
<td>800 kHz</td>
</tr>
<tr>
<td>5.7 µH</td>
<td>4.4 nF</td>
<td>7.2 µH</td>
<td>3.6 nF</td>
<td>1 MHz</td>
</tr>
<tr>
<td>5.7 µH</td>
<td>4.4 nF</td>
<td>24.2 µH</td>
<td>1.1 nF</td>
<td>1 MHz</td>
</tr>
<tr>
<td>5.7 µH</td>
<td>4.4 nF</td>
<td>15.5 µH</td>
<td>1.6 nF</td>
<td>1 MHz</td>
</tr>
</tbody>
</table>
The gate drive signals for these operating frequencies are shown in Figure 5. 41. It was found that the converter worked well for operating frequencies under 1 MHz.

![Figure 5. 41: Gate drive signal for full bridge converter at 100 kHz, 800 kHz and 1 MHz](image)

The full-bridge was further tested as 800 kHz, and the output of the primary converter and the output voltage of the pick up circuit is given in Figure 5. 42.

![Figure 5. 42: Output of the primary converter and pick up circuit at 800 kHz](image)

Power was delivered to the pickup circuit was measured and plotted against the primary track current in Figure 5. 43. The maximum power delivered was 83.6 mW.
5.6.3. Comparison of Push Pull and Full Bridge Converters

The power delivery capability of the two converters were compared against the primary track current and plotter in Figure 5.44.
The push pull converter evidently delivers significantly higher power levels than the full bridge converter. It must be noted that the push pull converter was operated around 3.5 MHz while the full bridge converter was operated at a lower frequency of 800 kHz. However, the operating frequency of the push pull converter is highly unstable as it is driven at full resonance. This makes it difficult to tune the secondary side and obtain power at a stable frequency. This becomes problematic when the control signal communication technique, discussed in Chapter 6 is implemented. Therefore, both converters have pros and cons, where one delivers more power and fully charges the energy buffer and the other has a stable operating frequency which enables wireless control signal communication. The full bridge converter was chosen for further development in wireless control of the microactuator as discussed in Chapter 6.

5.6.4. Pick Up Circuit Performance

The IPT primary side successfully delivered power to the pick up circuit, which was then rectified to charge the supercapacitor. The rectified DC output level varied depending on the AC pick up voltage and a rectified output of 15.8 V was achieved to successfully charge the supercapacitor bank. It can be seen that a DC output of 14.09 V was achieved at a peak to peak pick up voltage of 35.6 V. However, with a pick up voltage of 24.7 V peak to peak only produced a DC output of 4.80 V. Figure 5.45 shows several DC outputs produced by different pick up voltages.
Figure 5.45: Rectified output voltages for different pick up voltages
5.6.5. **Supercapacitor Energy Buffer Performance**

The rectified DC output was then fed to the supercapacitor energy buffer, which could be charged up to 15.87 V. Figure 5. 46 shows the charging curve of the supercapacitor which charges up to 12.2 V, as the oscilloscope is not capable of recording the full charge cycle of the supercapacitor in a single screenshot. The total energy stored in the energy buffer was 9.19 J, when charged to 15.87 V. The total charge time was approximately 15 minutes.

![Figure 5. 46: Supercapacitor charging curve](image)

The discharge profile of the supercapacitor was difficult to measure on the oscilloscope, as the minute energy draw after each actuation was not represented well on the screen.

5.7. **Performance Summary**

The performance of the wireless power supply can be summarized as seen in Table 5. 9. The IPT based wireless power supply successfully delivers 83.6 mW – 1.12 W to the
microactuator, which is then stored in the supercapacitor energy buffer. The energy buffer meets the 40 W peak power requirement during the 120 µs actuation.

Table 5.9: Summary of wireless power supply performance

<table>
<thead>
<tr>
<th>Primary Converter Topology</th>
<th>Push pull</th>
<th>Full Bridge</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power delivered</td>
<td>1.12 W</td>
<td>83.6 mW</td>
</tr>
<tr>
<td>Maximum primary current (rms)</td>
<td>1.09 A</td>
<td>0.82 A</td>
</tr>
<tr>
<td>Maximum output voltage in pick up (rms)</td>
<td>11.39 V</td>
<td>12.42 V</td>
</tr>
<tr>
<td>Maximum pick up current (rms)</td>
<td>253 mA</td>
<td>53.8 mA</td>
</tr>
<tr>
<td>Resonant operation on primary side</td>
<td>Always</td>
<td>Needs tuning</td>
</tr>
<tr>
<td>Resonant operation on secondary side</td>
<td>Needs tuning</td>
<td>Can be pre-tuned</td>
</tr>
<tr>
<td>Stable operating frequency</td>
<td>No</td>
<td>Yes</td>
</tr>
<tr>
<td>Equivalent capacitance of energy buffer</td>
<td>0.073 F</td>
<td></td>
</tr>
<tr>
<td>Maximum supercapacitor charge voltage</td>
<td>15.87 V</td>
<td></td>
</tr>
<tr>
<td>Energy stored</td>
<td>9.19 J</td>
<td></td>
</tr>
<tr>
<td>Actuations allowable</td>
<td>3600</td>
<td></td>
</tr>
<tr>
<td>Energy buffer charge time</td>
<td>15 minutes</td>
<td></td>
</tr>
</tbody>
</table>
Delivering the actuation command from the ABL platform to the microactuator wirelessly is the last step in making the actuator entirely wire-free. In the previously utilised solenoid microvalve the wired power was simply discontinued to return the valve to its default, open position. With the unique power supply arrangements of the novel microactuator, a separate actuation control signal needs to be generated. This introduced the additional challenge of incorporating wireless communication into the actuator module.

The control signal issued by the ABL platform is to toggle the present state of the valve. Once this toggle signal is received by the actuator, the micro-switch based position feedback system is used to determine the present position of the valve. Then a current is driven through the actuator to change the actuator to the opposite position.

Existing wireless communication methods, with lower power consumption and smaller real estate requirements were investigated. After weighing out the pros and cons, a control signal delivery strategy was designed using the existing IPT power link, of which the details will be discussed in this chapter.
6.1. Short Range Wireless Communication Technologies

Short range, low power wireless communication technologies with smaller real estate requirements, such as infra-red (IR), Bluetooth, radio frequency identification (RFID), near-field communication (NFC), ZigBee etc. were studied for feasibility.

IR communication is based on light waves just beyond the wavelength of visible red light, with frequencies varying between 430 THz – 300 GHz [119], but signal communication happens around the 38 kHz range. It is widely used in night vision, thermal imaging, remote controls for consumer electronics etc. The IR LED emits IR waves focused into a narrow beam. The receiver converts the IR signal to an electrical signal to decode the information. Line of site is very important, and it cannot penetrate through thick barriers such as walls. Although the maximum range is 10 m, a low power IR standard is also defined with a much shorter range of 20 cm [119]. The trade-off between communication range and power has made way for longer battery life for portable applications.

Bluetooth is a standard protocol used in short range radio communication between paired devices, such as mobile phones, computers, entertainment systems etc. It is suitable for voice and data transmission, with a maximum range of 10 m and data transfer rate of 2 Mbps, operating in the 2.45 GHz frequency band [119, 120]. Direct line of site is not required in Bluetooth which is advantageous over IR, for the microactuator application. Bluetooth low energy (BLE) is a later enhancement of the classic Bluetooth protocol to support low power applications with lower data rate requirements [120]. The power and peak current consumption is significantly lower with much shorter setup times.

RFID uses electromagnetic induction for communication purposes, mainly for identification, location, tracking and inventory applications. The reader transmits a high power RF signal to energise the passive tags to read data stored in their memory. RFID tags are small, flat and inexpensive and are designed to be attached to anything that require tracking or identification. RFID works at the 125 - 140 kHz, 13.56 MHz and 400 -
902 MHz or 2.45 GHz, and 4.8 or 5.8 GHz respectively in the low frequency (LF), high frequency (HF), ultra-high frequency (UHF) and super high frequency (SHF) bands [121]. The communication range can go up 8 m depending on the operating frequency [121].

NFC is a sub-category of RFID, which is capable of two-way communication. The initiator and target are both equipped with loop antennas within each other’s field to form an air-core transformer [122]. NFC communication can be active or passive, based on whether the target is powered or not. In passive operation, the target draws power to operate from the initiator. The initiator provides a RF carrier field, while the target device responds by modulating this carrier [122]. Passive targets can be tags, cards, stickers etc. If both initiator and target are powered active mode, NFC peer to peer communication is possible [122]. With the target powered, both initiator and target generate their own field alternately [122]. NFC operates in the globally available, unlicensed frequency of 13.56 MHz, with data rates varying between 106 - 424 kbps. Although the theoretical communication range is a maximum of 20 cm, the practical range is up to 10 cm [122]. Main advantages of NFC are the low power operation and not needing the receiver and initiator to be paired.

Zigbee is a low-power, low data rate and close proximity wireless communication protocol. Zigbee modules are used for home automation, medical device data collection, wireless light switches, electrical meters with in-home-displays, traffic management systems, and other low-power low-bandwidth applications requiring wireless communication. It is inexpensive in comparison with Bluetooth. The transmission range varies between 10 – 100 m, with line of site, but can be mesh networked to transmit data over long distance [123]. The data rate is 254 kbps and therefore best suited for transmitting data from a sensor or input device [123].

Table 6. 1 compares the main aspects of the short range wireless communication technologies currently available for the microactuator application.
### Table 6.1: Comparison of short range wireless communication technologies [119-123]

<table>
<thead>
<tr>
<th></th>
<th>Range</th>
<th>Frequency</th>
<th>Data rate</th>
<th>Energy / Power / Peak current consumption</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>IR</strong></td>
<td>0.2 – 10 m</td>
<td>300 GHz – 430 THz</td>
<td>115 kbps – 4 Mbps</td>
<td>48.2 μJ/b, 10.2 mA</td>
</tr>
<tr>
<td><strong>Bluetooth</strong></td>
<td>100 m</td>
<td>2.4 – 2.5 GHz</td>
<td>2.1 Mbps</td>
<td>1 W, &lt; 30 mA</td>
</tr>
<tr>
<td><strong>BLE</strong></td>
<td>50 – 150 m</td>
<td>2.4 – 2.5 GHz</td>
<td>1 Mbps</td>
<td>0.153 μJ/b, 0.01 – 0.5 W, 12.5 mA</td>
</tr>
<tr>
<td><strong>RFID</strong></td>
<td>20 – 100 cm</td>
<td>LF: 125 – 140 kHz</td>
<td>4 – 8 kbps</td>
<td>6.7 – 848 kbps</td>
</tr>
<tr>
<td></td>
<td>0.9 – 2.5 m</td>
<td>HF: 13.56 MHz</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>0.6 – 8 m</td>
<td>UHF: 400 – 900 MHz, 2.45 GHz</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>0.6 – 8 m</td>
<td>SHF: 4.8 GHz, 5.8 GHz</td>
<td></td>
<td></td>
</tr>
<tr>
<td><strong>NFC</strong></td>
<td>10 – 20 cm</td>
<td>13.56 MHz</td>
<td>106, 212, 424 kbps</td>
<td>50 mA</td>
</tr>
<tr>
<td><strong>Zigbee</strong></td>
<td>10 – 100 m</td>
<td>2.4 GHz</td>
<td>254 kbps</td>
<td>185.9 μJ</td>
</tr>
</tbody>
</table>

While NFC, BLE and Zigbee are potential wireless communication options for delivering the control signal to the actuator, they consume additional energy and require precious PCB real estate. The data communication strategy behind NFC and other RFID options work on the same principals of magnetic induction, and therefore is very similar to the IPT wireless power delivery. The control signal of the actuator is a simple binary, toggle signal which doesn’t need a complex communication protocol such as Bluetooth, IR or Zigbee. Therefore, an experimental communication technique was developed, where the existing IPT power transfer resources would be utilised to deliver the control signal.

### 6.2. IPT Based Wireless Communication

The binary control signal is issued by the ABL platform when each valve needs to be toggled. This signal is used to change the IPT operating frequency of the primary coil, where this variation is detected by the secondary pick up and actuation circuit. Then an
actuation current is delivered to the actuator in the desired direction to change the
current state of the valve. Figure 6.1 shows an overview of the communication technique
developed.

![Figure 6.1: Basic block diagram of control signal communication](image)

Once the ABL platform requests the valve’s state be toggled, this is communicated to the
FPGA platform controlling the switching of the IPT power converter. The FPGA then
alters the switching frequency of the FPGA for a short period of time, which can be varied
between few seconds to few minutes. It is important to allow sufficient time for this
frequency variation to be detected and acted upon by the secondary side. The pickup
circuit is equipped with an additional frequency detector to identify this change in the
IPT operating frequency to generate a high signal to control the power output to the
actuator. The output of the energy buffer is left disconnected normally, while a high
signal generated by the frequency detector is used to turn the power on for the actuator
control circuit. This method allows energy conservation in the system, while controlling
the actuation accordingly. Once power is supplied to the actuator, a position detection
signal is triggered and based on the present position, a current is driven through the
actuator coil to alter the magnetic field and change the state of the valve.

### 6.2.1. Frequency Detection Techniques

Several frequency detector techniques were investigated to deduce the best option for the
microactuator, the most important aspects being power consumption, PCB real estate
and accuracy. Analog filters, microcontroller based frequency counters and frequency to voltage converters were the main options that were investigated.

### 6.2.1.1. Microcontroller Driven Frequency Counter

Microcontroller based frequency counters are commonly utilised in frequency detection applications, due to straightforward implementation and high precision. The number of rising or falling edges within a fixed period, often set to 1 s, is accumulated to give the frequency of the respective signal. The maximum measurable frequency is limited by the maximum oscillation frequency of the microcontroller. Using this technique, a change in the IPT frequency can easily be detected, and the microcontroller can be utilised to generate the control signal to enable power to the actuator control circuit. The microcontroller based frequency detection method is illustrated in Figure 6.2.

![Figure 6.2: Microcontroller driven frequency detector](image)

The voltage generated at the IPT pickup is sinusoidal, which is converted to a square wave before counting. A very low power microcontroller which could operate with as little as 1.8 V consuming 30 μA/MHz, PIC16LF was chosen to minimize the power consumption. But low power microcontrollers such as this cannot drive the MOSFET switch directly due to output current limitations, requiring additional gate drivers. This increased the continuous power consumption of the overall system, which was not desirable.
6.2.1.2. Frequency to Voltage Converter

Another popular technique to detect the frequency is a frequency to voltage converter (FVC). A typical circuit as shown in Figure 6.3, which generates an output voltage proportional to the frequency of the input signal.

![Frequency to voltage converter circuit](image)

Figure 6.3: Frequency to voltage converter circuit

The input voltage $V_{in}$ with frequency $f_{in}$ is compared with the threshold voltage of the internal comparator. The falling edge of the input triggers a one-shot timer for a period of time defined by $1.1R_tC_T$. During this period, a current $i$ will flow to charge $C_L$ where the amount of charge $Q$ is fixed. This will generate an output voltage $V_{out}$ proportional to the frequency $f_{in}$, defined by (6.1).

$$V_{out} = f_{in} \left( \frac{R_L}{R_S} \right) (1.9)(1.1R_TC_T) \quad \text{(6.1)}$$

$R_S$ is variable to allow for a required scale factor of the output, so a gate drive voltage can be produced to drive the power control MOSFET when a desired frequency has been detected.

The main drawback with this technique too is the continuous power demand, to monitor the frequency of the IPT system.
6.2.1.3. Analog Passive Bandpass Filter

Using an analog bandpass filter, either passive or active, is another option for identifying the desired change in frequency. Passive filters are composed purely of passive components such as resistors, capacitors and inductors with no continuous power requirement. They provide guaranteed stability, scale better to large signals and are cheaper to construct [124]. Active filters require continuous operational power due to the use of transistors or opamps but uses no bulky inductors and are easier to design and tune [124]. They produce high gains and are generally more compact [124].

The analog filter is designed to continuously monitor the IPT power input signal and generate a high output when the frequency changes to desired control frequency $f_{\text{control}}$. The passband of the filter is designed with its centre frequency at $f_c$, to generate a higher gain at this frequency. During the regular power transfer operation at frequency $f_p$, the filter output would be attenuated as it falls into the stopband of the filter. When the IPT frequency changes to $f_c$, the high output of the filter is used to control the actuation of the valve, by controlling the current flow into the actuator control circuit. The expected frequency response of the filter is shown in Figure 6.4.

![Figure 6.4: Frequency response of the filter](image)

An analog passive filter was preferred in this context as continuous operational power is not required, as opposed to an active filter which would draw power continuously.
6.2.2. Control Signal Detector Design

When designing the control frequency identification circuit, the first step was to identify the most suitable bandpass filter topology, which was then followed by the filter circuit design. The filter response for the designed circuit was simulated, implemented and tested. The design was then extended for several valves operating at different control frequencies on the same microfluidic platform.

6.2.2.1. Filter Type Selection

There are several popular filter design topologies such as Butterworth, Chebyshev, Bessel and Elliptic filter with different key performance parameters. As no filter topology is ideal, each topology has trade-offs for optimum performance to meet the requirements of the context. Table 6.2 compares the basic characteristics of these four main filter types.

<table>
<thead>
<tr>
<th>Passband</th>
<th>Butterworth</th>
<th>Chebyshev</th>
<th>Bessel</th>
<th>Elliptic</th>
</tr>
</thead>
<tbody>
<tr>
<td>Stop band</td>
<td>Maximally flat</td>
<td>Ripples</td>
<td>Flat</td>
<td>Ripples</td>
</tr>
<tr>
<td></td>
<td>Good attenuation</td>
<td>High attenuation</td>
<td>Slow initial attenuation</td>
<td>Ripples</td>
</tr>
<tr>
<td>Phase response</td>
<td>Moderate distortion</td>
<td>Large phase shifts near cut-off frequency</td>
<td>Linear phase shift</td>
<td>Highly non-linear phase shift</td>
</tr>
<tr>
<td>Transition band</td>
<td>Slow</td>
<td>Sharp</td>
<td>Slow</td>
<td>Sharpest</td>
</tr>
</tbody>
</table>

Butterworth filters provide the best passband response which is the flattest of all, but the trade-off is the slower roll off in the transition band [125]. Chebyshev filters on the other hand provide a much steeper roll off in the transition region, but with ripples in the pass or stop bands [125]. The Bessel filter is optimized for the best phase response, which in this case has minimal importance [125].

The linearity of the phase response has negligible effect in the microactuator context as the key focus is generating a sufficiently large output for the control frequency and better stopband performance. As the microactuator has limited space, the component count of the filter is vital. Therefore, a lower order Chebyshev filter with a sharper roll off in the transition band as more feasible than a Butterworth filter with a similar roll off, which would have to be of a much higher order.

### 6.2.2.2. Actuator Control Circuit Design

To maintain a lower component count and hence minimize the PCB real estate requirements, a Chebyshev filter with the specifications outlined in Table 6.3 was designed for the microactuator.

<table>
<thead>
<tr>
<th>Filter type</th>
<th>Bandpass</th>
</tr>
</thead>
<tbody>
<tr>
<td>Centre frequency</td>
<td>1 MHz</td>
</tr>
<tr>
<td>Passband bandwidth</td>
<td>80 kHz</td>
</tr>
<tr>
<td>Passband ripple</td>
<td>0.1 dB</td>
</tr>
<tr>
<td>Stopband attenuation</td>
<td>-45 dB</td>
</tr>
</tbody>
</table>

A third order filter was designed, with its output voltage rectified to directly drive the power control MOSFET switch, as shown in Figure 6.5.
The simulated magnitude and phase response of the designed filter is shown in Figure 6. 6, over the range of 10 kHz – 100 MHz. It can be clearly seen that the filter attenuates the IPT power transfer frequency and does not generate a high output to drive current into the actuator at this frequency. When the 1MHz communication frequency is detected by this filter a control signal is generated to drive the actuator circuit to open/close the valve.

6.2.2.3. **Wireless Control Implementation**

The implemented control signal detector circuit is shown in Figure 6. 7. The filter circuit is not directly connected to the IPT pickup coil, as it would alter the resonant frequency of both filter circuit and the power pickup circuit. Instead the filter circuit is placed in the close vicinity of the IPT pickup, in order to resonate when the pick up coil experiences the control command at the 1MHz frequency. This ensures that the IPT circuit transfers power independently to the energy buffer and when the frequency is changed. This indicates a command to toggle the state of the valve the filter circuit delivers a DC signal.
to turn on the power control MOSFET which otherwise cuts the power to the remainder of the actuator circuit. When power is enabled to the actuator control circuit, the valve position detector circuit feeds back the current position to the control circuit. Based on whether the valve is currently in the open or closed position, the logic circuit generates a control pulse to drive a current through the coil, which toggles the position of the valve. A PMZ200UNE MOSFET has been chosen for power flow control, which is capable of very fast switching, low threshold voltage and has very small dimensions of $1.0 \times 0.6 \times 0.48$ mm.

![ Implemented control signal communication module ]

6.2.3. Performance Analysis of Communication Module

The implemented communication technique was tested for robustness and reliability. The filter circuit was tested at a variety of frequencies to obtain the frequency response, as shown in Figure 6. 8. It can be seen the centre frequency is at 1 MHz as desired and exactly the same as in the simulated filter response shown in Figure 6. 6. Any signals outside the passband are attenuated to avoid the generation of a control signal to enable power to the actuator control circuit.
Figure 6.8: Frequency response of bandpass filter with centre frequency of 1 MHz

The IPT converter was driven at its power transfer frequency of 800 kHz and the output of the filter was monitored, as shown in Figure 6.9.

Figure 6.9: Filter output at IPT power transfer frequency
The low DC output ensures the power control MOSFET stays turned off to keep the actuator control circuit turned off.

The IPT converter was then driven at the control signal communication frequency of 1 MHz and the output of the filter was monitored, as shown in Figure 6. 10 (a) and (b).

![AC output of filter](image1) ![DC output of filter](image2)

(a) AC output of filter  
(b) DC output of filter

*Figure 6. 10: Filter output at IPT power transfer frequency of 800 kHz*

The high DC output turns the power control MOSFET on to power up the actuator control circuit, which then drives the actuator to toggle state of the valve. And once the converter resumes the power transfer frequency, the MOSFET is turned off and actuator control circuit is shut down.
Chapter 7

Overall Performance of Bi-Stable Wireless Microactuator

7.1. Full System Operation

The individual modules of the actuator discussed in the previous chapters are integrated to form the complete actuator and wireless power transfer system. The IPT full bridge converter is driven directly using the DC power supply, with the control handled by the FPGA onboard the NI 9401 PI controller. The MOSFET switches in the IPT converter are driven by the PI controller at the desired operating frequency, using PWM waves. During this entire time power is transferred to the supercapacitor energy buffer to keep it fully charged.

The PI controller is configured to communicate with the ABL platform and receive the valve toggle command. The PI controller then changes the switching frequency of the IPT converter to 1 MHz, which is the predetermined communication frequency. This frequency variation lasts for approximately 1 minute allowing sufficient time to complete the actuation, before the converter reverts to the power transfer frequency. The valve control process is outlined in Figure 7.1.
Once the ABL platform issues the valve toggle command, the PI controller changes the operating frequency of the IPT converter by switching at 1 MHz. This frequency change generates the secondary current on the IPT pickup at 1 MHz. The bandpass filter circuit with 1 MHz centre frequency is used for frequency detection, which generates a high output when the frequency is within the passband, specifically at 1 MHz. The output of the filter is used as the gate control of the MOSFET switch that enables and disables power flow into the actuator control circuit. When the power flow control switch is closed, the output of the supercapacitor is connected to the actuation circuit. The actuator position detection circuit is powered up at this point, and if the micro-switch is compressed a short trigger pulse is given one of the logic outputs, and if the switch is uncompressed the other logic output generates a trigger pulse. The generated trigger pulse drives a timed a current pulse through the EP magnet coil in the desired direction to open or close the valve. The IPT operating frequency is maintained at 1 MHz for approximately one minute, which allows time for the signal to transmit through the circuitry and perform the actuation.

7.2. Summary of Microactuator Performance

In summary, the main features of the developed microactuator include:

- A novel bi-stable microactuator mechanism based on electropermanent magnets
- Ultra-low energy consumption per actuation and zero energy consumption between actuations
- IPT based wireless power supply for wire-free operation
- Self-sustained operation via energy buffering
- Wireless actuator control over IPT link using frequency variation technique
- Significant deflection range and force capability
- Super-fast micro second actuation

The specific operation parameters of the microactuator are presented in Table 7.1.

<table>
<thead>
<tr>
<th>Actuation technology</th>
<th>Electropermanent magnet</th>
</tr>
</thead>
<tbody>
<tr>
<td>Supply voltage</td>
<td>15.75 V</td>
</tr>
<tr>
<td>Magnetising Coil current</td>
<td>2.5 A</td>
</tr>
<tr>
<td>Actuation speed</td>
<td>120 μs</td>
</tr>
<tr>
<td>Minimum actuation energy</td>
<td>0.9 mJ</td>
</tr>
<tr>
<td>Stable actuation energy requirement</td>
<td>2.4 mJ</td>
</tr>
<tr>
<td>Power consumption (during actuation)</td>
<td>40 W</td>
</tr>
<tr>
<td>Power consumption (in between actuations)</td>
<td>0 W</td>
</tr>
<tr>
<td>Maximum holding force</td>
<td>220 mN</td>
</tr>
<tr>
<td>Maximum attraction force</td>
<td>98 mN</td>
</tr>
<tr>
<td>Maximum deflection</td>
<td>2.5 mm</td>
</tr>
<tr>
<td>Wireless power topology</td>
<td>IPT</td>
</tr>
<tr>
<td>Wireless power delivery capability</td>
<td>1.12 W</td>
</tr>
<tr>
<td>Energy buffer technology</td>
<td>Supercapacitor</td>
</tr>
<tr>
<td>Energy storage capability</td>
<td>9.19 J</td>
</tr>
<tr>
<td>Maximum actuations allowable</td>
<td>3600</td>
</tr>
<tr>
<td>Energy buffer charge time</td>
<td>15 minutes</td>
</tr>
</tbody>
</table>
7.3. Performance Comparison with Existing Microvalve Actuators

The performance of the novel microactuator was compared with existing microvalves at different levels. First it was compared with the solenoid valves used on the ABL platform for power and energy consumption. Then the power and energy requirement of the novel valve and the existing solenoid valve was compared when it was operated over a full experimental run of the ABL platform. Next the actuator was compared with the main active microvalve types for power and energy consumption. It was finally compared with state-of-the-art bi-stable microvalves discussed in Chapter 2.

7.3.1. Comparison with Normally Open Solenoid Pinch Valve

The performance of the actuator was compared with the solenoid pinch valve on the ABL platform, described in section 2.2, which the novel valve is intended to replace. In order to compare the actuators, a scenario where each valve was opened for 5 minutes and closed for 5 minutes. The power and the cumulative energy consumption for each actuator were plotted in Figure 7.2. As the cumulative energy consumption of the EP magnet actuator is 44 mJ, it has been represented in mJ on the secondary axis, while the solenoid actuator energy which accumulates to 4.35 kJ, is plotted in Joules, for better visualisation.

The target microfluidic platform [21] for which this microvalve was designed currently uses 4 solenoid microvalves for an experimental run which last 3 days, consuming 3 W of power when the valves remain closed. As each set of valves toggle states when the experiment enters a new stage and is held in that position for prolonged period of time, the energy consumption and heating of the system is significantly high. Therefore, the new actuator provides a much more energy efficient setup. The total energy consumption profile of the current platform was compared with the same platform utilizing the novel microvalve actuator as shown in Figure 7.3. Energy consumption of the current system,
shown in blue, is significantly higher than the energy consumption with the new actuator shown in red. The power consumption with the novel actuator shows spikes at the points where the valves require actuation and zero otherwise.

**Figure 7.2**: Novel actuator performance comparison with solenoid valve actuator in ABL platform

**Figure 7.3**: Power and energy performance of EP magnet actuator vs current solenoid valves in target microfluidic platform
7.3.2. Comparison with Other Active Microvalves

To obtain a perspective on how the novel actuator performs in comparison with other active microvalve types, such as piezoelectric, pneumatic, electromagnetic and SMA valves, their power and energy consumption was compared. As the power and energy consumption of the designed actuator is strictly dependent on the number of actuations, this comparison is based on the valves operated over one hour, where each type of valve is opened every 10 minutes and left open for 5 minutes. Based on power and energy consumption data available about commercial and research level microvalves of different types [28, 52, 56, 64, 126-128], the average power and total energy consumption range of each valve type was plotted in Figure 7.4.

To justify the lowered power consumption of the EP magnet actuator, the average power consumption over an hour was taken, which falls within the same range of other electromagnetic actuators. It is also evident that the average power consumption is significantly lower than some piezoelectric and pneumatic actuators. In comparison with other active microvalve types the energy consumption over one hour is significantly lower. While electromagnetic and SMA valves consume several kilo-Joules of energy, piezoelectric and pneumatic valves consume several hundreds of kilo-Joules in an hour. The EP magnet actuator consumes an average of 24 mJ in one hour for the aforementioned operating conditions. The energy consumption is plotted on a logarithmic scale, for clearer representation of the EP magnet. It is evident that the EP magnet actuator has significant reduced energy requirements, when compared with all other active valve types.
7.3.3. Comparison with Bi-Stable Microvalves

While the general comparison above gives insight into the significance of this novel design, the EP magnet actuator was further compared with some state-of-the-art bistable valves, as shown in Table 7.2. There were very few comparable valves to be found in literature, which offer functionalities of similar nature. The same valves discussed in section 2.4 were used for comparison with the EP magnet actuator. The EP magnet actuator claims the lowest energy per actuation, fastest actuation and highest deflection. The force capability of the novel actuator is similar to the work of Megnin, C. et al. (2012), but provides the best return on the amount of energy consumed. The only drawback is the high momentary power, supply voltage and current requirements, which is much higher than those valves in comparison. However, this is managed well with the integration of the supercapacitor energy buffer. Additionally, the new bi-state actuator provides wireless power and control capability which is not found in any of the valves in comparison, or any other valve available to date. This makes the new microvalve actuator absolutely sought after in microfluidic applications.

Figure 7.4: Power and energy consumption of EP magnet actuated valve in comparison with other externally actuated microvalves
Table 7.2: Comparison of EP magnet actuator and state-of-the-art bistable valves

<table>
<thead>
<tr>
<th>Microvalve</th>
<th>Current (A) / Voltage (V)</th>
<th>Power</th>
<th>Pulse</th>
<th>Actuation Energy</th>
<th>Force</th>
<th>Deflection</th>
</tr>
</thead>
<tbody>
<tr>
<td>Solenoid Solutions Inc.</td>
<td>-</td>
<td>0.65-9W</td>
<td>20-50ms</td>
<td>13-450mJ</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Sutanto, J. et al. (2007)</td>
<td>570mA &lt; 5V</td>
<td>2.85W</td>
<td>0.5ms</td>
<td>1.425mJ</td>
<td>-</td>
<td>20-35µm</td>
</tr>
<tr>
<td>Bohm, S. et al. (2000)</td>
<td>500mA -</td>
<td>100ms</td>
<td>-</td>
<td>2.15-2.7N</td>
<td>200µm</td>
<td></td>
</tr>
<tr>
<td>Capanu, M. et al. (2000)</td>
<td>340mA 1.38V</td>
<td>-</td>
<td>5.3ms</td>
<td>2.48-4.26mJ</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Chang, P. J. et al. (2012)</td>
<td>3.8mA</td>
<td>50mW</td>
<td>-</td>
<td>10-15nN</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Megnin, C. et al. (2012)</td>
<td>-</td>
<td>500mW</td>
<td>200ms</td>
<td>100mJ</td>
<td>290mN</td>
<td>50µm</td>
</tr>
<tr>
<td>EP magnet valve (2017)</td>
<td>2.5A 15V</td>
<td>40W</td>
<td>0.12ms</td>
<td>0.97-2.5mJ</td>
<td>220mN</td>
<td>2.5mm</td>
</tr>
</tbody>
</table>

In summary, the experimental results have shown that 0.97-2.5 mJ of energy is sufficient to actuate the valve, with a current pulse varying between 44-150 µs. The optimum number of turns was found to be 200, theoretically and experimentally. The actuator provides a large deflection range varying from 0.3-2.5 mm. It provides a maximum holding force of 220 mN and attraction force as large as 98 mN. A 15 V, 2.5 A pulse lasting 120 µs is used to actuate the microactuator which leads to a momentary power consumption of 40 W. However, with the zero energy consumption between actuations the energy consumption of this actuator is much lower than any state-of-the-art bistable valves.
Chapter 8

Conclusions and Future Recommendations

8.1. General Conclusions

This thesis contains research and development work conducted on a novel type of microactuator suitable for microfluidic valves. The focus of this work was on delivering the following features as the major outcomes of this PhD project:

- Development of a novel microactuation mechanism, which is highly energy efficient in comparison with existing microfluidic valve actuators.
- Design and implementation of an IPT based wireless power transfer system and energy buffer, for self-sustained, safe and flexible operation of the microactuator.
- Development of a wireless control of the microactuator with the aid of existing electronics to avoid additional wireless communication modules.

Chapter 1 gives a brief introduction to microfluidic systems, such as their functionality, advantages they have over macro-level fluidic analysis, and some of the key components used. The functionality of the automated bio-chemical laboratory, which is a microfluidic platform developed to screen the effects of toxins on Zebra fish embryos is discussed, while focusing on the role of the on-board microvalves. It has been shown that the solenoid valves used in this system, as with all other solenoid valves, are large consumers of electric power. This makes it very difficult to power such systems with smaller batteries, for portability. A secondary effect of the large power consumption is the significant heating in the system, which harms the embryos and changes chemical properties. The wired power connections of the valves are hazardous, electrically unsafe in a bio-fluidic environment, prone to collecting bacteria and other contaminants, chaotic
and mechanically restrictive. This leads to the need for a novel microvalve actuator, which has energy efficient actuation and wirelessly powered. The chapter also gives an overview of wireless power, its common applications, common WPT technologies in use, benefits offered to the microfluidic platform with the introduction of wireless power.

A brief overview of different active microvalve types is presented at the beginning of Chapter 2. Active microvalves types with external microactuators are focused on to identify their power requirements, and electromagnetic valves were found to be the highest power consumers. Operating principles of various electromagnetic valves and the physical parameters governing the performance of these valves were studied further to identify causes of the large power requirement. Solenoid valves, which are utilised in the ABL platform, are a subset of electromagnetic valves are driven by a current carrying and require continuous power to operate. Novel state-of-the-art valves with low power actuation, bi-stable functionality or wireless power were investigated for feasibility in the ABL platform, to improve its power performance. Studying these valves have shown that while some valves feature one or two of these aforementioned attributes, a single valve which delivers wireless power, low energy actuation and bi-stable operations does not exist in current literature.

Chapter 3 then presents an overview of the proposed novel actuator, which features an ultra-energy efficient actuation mechanism, coupled together with wireless power and control to drive microfluidic valves. The proposed actuation mechanism is based on electropermanent magnets, which consumes energy for a fraction of a second during actuation and consumes zero energy in between actuations. This facilitates bi-stable operation, a highly desirable feature in microactuators. Wireless power is enabled through inductive power transfer, which is coupled with supercapacitor energy buffering to meet the peak power requirement of the actuator and self-sustainability. A wireless control technique is also proposed for the microactuator, to deliver the binary valve toggle command from the control platform to the actuator control circuit. The combined functionality of these modules is outlined, and each module is then introduced briefly to make way for further discussion in the following chapters.
The design details of the novel actuation mechanism proposed in this thesis is presented in Chapter 4. This novel actuator has been designed using electropermanent magnets. Electropermanent magnets combine a semi-hard and hard magnetic material with a current carrying coil. The actuation is governed by the coercivity and residual magnetism of the chosen magnetic materials. After studying the properties of magnetic materials, NdFeB was chosen to be the hard magnet due to its very high coercivity. The much lower coercivity of AlNiCo5 determined it to be the most suitable semi-hard magnetic material. When a sufficiently large current is applied to the coil briefly, it alters the magnetisation of the AlNiCo magnet while the NdFeB magnet’s magnetisation remains unaffected. Due to the similar residual magnetism of both materials, they generate similar magnetic flux when the external magnetic field generated by the coil current is removed. When the AlNiCo magnetic field aligns with the NdFeB an external magnetic field is generated to attract the plunger and open the valve. When a current in the opposite direction alters the AlNiCo magnetic field, the opposing magnetic fields cancel out to retract the plunger and close the valve. Theoretical analysis of this novel design showed that the magnetic materials should be chosen to have significantly different coercivity levels but very similar residual magnetism values for successful valve actuation. Theoretical analysis also highlighted that it is vital to have identical cross-sectional areas for each magnet to generate a similar flux concentration. FEMM simulations of the designed actuator proved the actuation concept to be viable and was therefore implemented.

The electrical control circuit designed and implemented, identifies the present state of the valve and toggles this state by driving a timed current pulse through the EP magnet coil. Experimental results showed the best power and energy performance could be achieved with a 200 turn coil. Stable actuation was achieved with a coil current of 2.5 A and a supply voltage of 15.75 V applied for 120 µs. The power consumption during actuation was 40 W, which lasts only 120 µs, leading to an energy consumption of 2.4 mJ. The actuator was physically capable of delivering a 220 mN holding force and 98 mN attraction force, and a maximum deflection of 2.5 mm.
Design and implementation details of the wireless power supply is discussed in Chapter 5. Existing wireless power supply technologies are summarised highlighting their advantages, disadvantages, power levels and power transfer range. IPT was chosen as the most suitable power transfer mechanism to power the developed microactuator, due to the maturity of the technology. A single loop IPT primary track was designed and magnetically coupled with a secondary circular coil of 7.2 mm diameter and inductance of 24.2 \( \mu \text{H} \). A parallel tuned, current-fed autonomous push pull converter operating at 3.5 MHz was initially used for the primary side. Due to frequency instabilities in this converter a series tuned, voltage-fed full bridge converter operating at 800 kHz was also investigated, as frequency stability is vital for wireless control, as described in Chapter 6. The IPT secondary circuit was designed with a series tuned pick up, with its rectified output charging a supercapacitor energy buffer. The constant voltage output of the series tuned pick up allowed for a stable charging voltage for the supercapacitors. Due to the peak power requirement of 40 W during actuation, energy buffering was essential. A series connected supercapacitor bank was designed to store 9.2 J on board. Experimental results showed that the wireless power supply using the autonomous push-pull converter was capable of delivering a maximum power of 1.12 W to the actuator at 3.5 MHz. The full bridge converter driven at 800 kHz was only capable of delivering a maximum power of 83.6 mW. The supercapacitor energy buffer takes 15 minutes to fully charge and is capable of sourcing 3600 actuations before having to recharge.

Chapter 6 investigates the wireless control signal communication for the microactuator to create a truly wireless actuator. The intention of adding wireless communication is to deliver the binary control signal from the main ABL platform controller to the microactuator. A summary of present wireless communication technologies suitable for the microactuator application is given initially. These techniques evidently are an overkill for such a simple task, and therefore a technique was developed to use the same IPT power transfer link for signal communication. When the ABL platform issues a command to toggle the actuator state, the operating frequency of the IPT primary converter is changed to 1 MHz for 1s. The PI controller driving the primary full bridge converter, changes the switching frequency of the gates to 1 MHz. A passive bandpass filter based,
frequency detection circuit identifies this change in frequency to generate a high output to turn the actuator control circuit on and drive the actuator. Once the actuator state has been toggled, after 1 s, the wireless power transfer resumes at 800 kHz. Experimental results showed successful detection of the wireless control command to drive the actuator.

Chapter 7 outlines the overall operation of the complete actuator. The actuation sequence is as follows. The actuation command issued by the ABL platform is given to the cRIO controller, and the operating frequency of the IPT converter is changed to 1MHz accordingly. The frequency detector on the microactuator identifies the change in frequency to open a switch to enable power to the actuator control circuit. This triggers a command to detect the current position of the actuator, and a current is driven through the EP magnet coil to toggle the state of the actuator. This process is completed in less than 1 second, after which the controller changes the operating frequency of the IPT converter back to 800 kHz. The power to the actuator is therefore disabled, by closing the switch, to conserve energy in the actuator.

The performance of the developed novel actuator was compared with the solenoid valves currently used on the ABL platform. It is shown that if both are actuated every 5 minutes to toggle their state over 1 hour, the cumulative energy consumption of the solenoid valve is 4.35 kJ, while the EP magnet actuator only uses 44 mJ. A significant reduction in energy consumption was also observed on the ABL platform during a full experimental run, with the designed actuator and the same solenoid valves. The average power consumption of the designed microactuator, with the valve is opened and closed every 5 minutes, was compared with other active types such as electromagnetic, piezoelectric, pneumatic and SMA microvalves. The average power consumption of the designed actuator was similar to the higher end of electromagnetic valves, but lower than most piezoelectric and pneumatic microvalves. However, the total energy consumption of the EP magnet actuator was significantly lower than all these valves. The EP magnet actuator was also with some of the comparable state of the art, bi-stable valves for power, energy and physical performance. The designed microactuator displayed the lowest energy
requirement per actuation of 0.97 – 2.5 mJ, when compared with valves developed by Capanu, M. et al. (2000), Sutanto, J. et al. (2007) etc. The deflection of 2.5 mm offered by the EP magnet actuator was higher than all others in comparison, which offer deflections in the micro-meter order. The designed microactuator also offers the fastest actuation of 120 µs, which is lower than all of the others including Sutanto, J. et al. (2007) which has the fastest actuation speed of 500 µs. The force capability of this microactuator is 220 mN, which is similar to that of Megnin, C. et al. (2012), while the work of Bohm, S. et al. (2000) have much higher forces in the Newton order. Overall these key features show that this microactuator has very high end state-of-the-art performance in a single package.

This ultra-energy efficient, bi-stable novel microactuator equipped with wireless power is successful in all aspects. It has given birth to a new class of microactuators for microfluidic applications. The versatility of this design allows for it to be adapted in any application requiring bi-stable actuation.

8.2. Thesis Contributions

As a whole, this thesis contributes a completely novel, wirelessly powered and controlled, ultra-energy efficient bi-stable microactuator suitable for microfluidic applications. The most unique feature of the developed microactuator is the electropermanent magnet based actuation mechanism, which displays a significant reduction in the actuation energy requirement. The lower energy requirement has led to the microactuator being wirelessly powered and controlled, which is a feature not seen among other microfluidic actuators found in literature.

The main contributions of this thesis include:

- A novel, ultra-energy efficient, bi-stable, ultra-fast 120 µs actuation mechanism based on electropermanent magnets, suitable to drive microfluidic valves, requiring 2.5 mJ per actuation, delivering forces up to 220 mN and a maximum deflection of 2.5mm.
• Two primary IPT power supplies operating at 800 kHz and 3.5 MHz to deliver 83.6 – 1.12 W of power to the microactuator wirelessly.

• A wireless power and control circuit to drive the microactuator, complete with a series tuned secondary IPT pick up absorbing up to 1.12 W and a supercapacitor energy buffer capable of storing 9.2 J and sourcing 3600 actuations.

• A unique, wireless actuation control mechanism, using frequency variation at 1MHz in the IPT module, to deliver the actuation signal from the microfluidic platform to the actuator control circuit.

8.3. Publications

One journal paper and five conference paper has been published as a result of this work:


Two more journal papers are intended to be published in future titled:


8.4. Recommendations for Future Work

This novel microactuator has tremendous potential in the microfluidics industry, as it delivers features which are not found in current state-of-the-art microvalves. Several future works are recommended to enhance the design and commercialise this microactuator.

1. Improvements in Electronic Circuit and Control Design
As the described work is a first prototype of the microactuator, the electronic design of the entire actuator has potential for further improvement. With the proof of concept design in place, future researchers can improve the converter specifications to deliver more power to the microactuator and charge the supercapacitor bank faster. The energy efficiency of the actuator has potential to be reduced beyond 2.5 mJ per actuation with the use of more optimal components. Further improvements in the IPT track and coil design can also allow better power transfer to the load.

2. Further Investigation and Analysis of Experimental Results

The experimental results discussed in Chapter 4, have left some unanswered questions about the variation in the required current to magnetise AlNiCo with less than 200 turns. Further experiments and analysis will need to be conducted to investigate as to why these variations exist.

3. Optimisation of Magnetic Structure Design

The current magnetic structure has been designed for easy implementation and fast prototyping. The dimensions and shape of both magnets need to be redesigned to find the optimal dimensions for a chosen valve enclosure which would facilitate better actuation and energy conservation. For example, for a cylindrical enclosure, disc shaped magnets can be investigated. With the aid of further FEMM analysis the dimensions of the hard and semi-hard magnets can be deduced for best performance for a given valve shape.

4. Improvements to Mechanical Design

The mechanical design in the current structure has a lot of room for enhancement. Firstly, a suspension mechanism needs to be designed for the plunger to hold it in place when it retracts from the EP magnet. At present, the plunger retracts purely as a result of gravity which could lead to undesired displacements. A non-
metallic spring suspension system would be a potential option to investigate in future.

The position feedback tact switches in the current design are manually attached with glue to the actuator pole pieces. This is quite delicate and only good as a proof of concept design. It is highly recommended to embed the micro switch into the pole piece to make the position feedback mechanism more precise, reliable and robust.

The plain magnetic rods in the EP magnet posed many challenges when winding the coil in an even manner. In future, for mass scale production grooved magnetic rods are recommended, which can better hold the coil wire in place.

A suitable valve enclosure is also recommended as a future requirement to place the components in a single package for further testing.

5. Miniaturisation

Although the final prototype was designed using SMD components, it has not been miniaturised to optimise the PCB real estate, as experimental work needed to be conducted on the current prototype. Therefore, it is recommended that a further miniaturised future design be implemented to fit the microactuator, control and wireless power supply within the valve enclosure.
References


[75] IBS Magnet, "Permanent magnets materials and magnet systems."


[77] Eclipse Magnetics, "Ferrite magnets/ceramic magnets datasheet."


