

Development of an online broadband impedance spectroscopy extraction system of PV modules through switch-mode converters



Prepared by:

Lindani Shelembe

Department of Electrical Engineering
University of Cape Town

25 July 2023

Submitted to the Faculty of Engineering and the Built Environment (EBE) at the University of Cape Town, in fulfilment of the academic requirements for a Doctor of Philosophy (Ph.D.).

Key Words: Photovoltaic, Impedance Spectroscopy, APRBS, Chirp, QFT, Wide Bandgap Semiconductors, Multiphase Converter.

The copyright of this thesis vests in the author. No quotation from it or information derived from it is to be published without full acknowledgement of the source. The thesis is to be used for private study or non-commercial research purposes only.

Published by the University of Cape Town (UCT) in terms of the non-exclusive license granted to UCT by the author.

Declaration

- I know that plagiarism is wrong. Plagiarism is using another's work and pretend that it is one's own.
- I have used the IEEE convention for citation and referencing. Each contribution to, and quotation in, this final year project report from the work(s) of other people, has been attributed and has been cited and referenced.
- This report is my own work.
- I have not allowed and will not allow anyone to copy my work to passing it off as their own work.

Name: Lindani Shelembe

Signature:

Signed by candidate

Date: 25 July 2023

Abstract

Solar PV systems have traditionally relied on terminal DC current and voltage characteristics for online diagnostics. This approach monitors the current and voltage levels of a module, and a change in these levels indicates a system anomaly. However, this information only alerts to a decrease in the current or voltage levels, but not the possible cause(s) for the anomaly. To enable more detailed online diagnostics, linear impedance spectroscopy, a frequency-based characterization technique implemented through a DC-DC switch-mode converter, has been shown to provide fast and reliable impedance information that can be utilized to diagnose the condition of batteries and fuel cells in real-time. However, the implementation of this technique in a PV system presents challenges that have not been identified or addressed in current research.

To develop such a system for a solar PV module, an excitation source with wide loop bandwidth capabilities is required to account for the broader impedance bandwidth of a PV module compared to batteries or fuel cells. The excitation signals must be designed such that they can minimize the nonlinearities observed at lower and higher frequencies, which result from induced excitation produced by the nonlinear DC-DC switch-mode converter. Additionally, a robust control system with variable-point measurement capabilities is required to ensure that when the operating point of the PV system changes, measurements can be taken in a linear manner, thus ensuring compliance with linear measurements.

This study presents an advancement to switch-mode converter-based AC impedance extraction for PV modules by proposing and designing a wide-bandwidth multiphase switch-mode DC-DC converter-based system to perform module-level online impedance measurements on the solar module. The system will be designed to allow a solar PV module to provide regulated power to a DC load, while extracting its AC impedance, revealing detailed information about the module's internal changes and performances, thus improving diagnostic capability. The multiphase converter was designed using wide bandgap Silicon Carbide semiconductor switches to improve switching utilization at 500kHz switching frequency, with lead control compensation in the current loop, to enable a wider loop bandwidth for the converter. Amplitude modulation is adopted in both Chirp and Pseudorandom excitation signals to mitigate frequency nonlinearities. Lastly, Quantitative Feedback Control Theory is implemented to obtain accurate impedance measurements for varying operating point conditions – which are typical for solar modules.

The impedance extraction system results showed a limitation in the design of the loop bandwidth; the most practically realizable loop bandwidth was 10% of the switching frequency. However, signal analysis showed that with well-designed amplitude modulation, the attenuation at high excitation frequencies could be mitigated in both digital and analog excitation signals, albeit at the cost of increased system complexity. With variable-point control, results showed that the Quantitative Feedback Control can keep the induced excitation signals within the required specifications during variable load and operational conditions for better impedance measurements. Impedance measurement results,

subsequently, showed good error performance, when benchmarked against a commercial frequency response analyser for partial shading, hot spots, microcracks, and variable-point tests. The maximum percentage impedance error observed was 3.36% – observed at low frequencies under partial shading – and was within acceptable margins when compared to existing research. Thus, the proposed system provides a satisfactory advancement to converter-based linear impedance spectroscopy for solar PV modules.

Table of Contents

Abstract.....	iii
List of Figures.....	xii
List of tables.....	i
List of abbreviations.....	ii
Chapter 1 Introduction	1
1.1 Background to the study.....	1
1.2 Problem statement.....	1
1.2.1 The high impedance bandwidth of solar PV modules.....	3
1.2.2 Linearity of the excitation signals under induced excitation in high bandwidth systems..	3
1.2.3 Control of online impedance extraction under variable-point conditions	4
1.3 Research questions	5
1.4 Purpose and contributions of this study	5
1.4.1 A wide-bandwidth multiphase boost excitation source for a solar module to perform high-frequency impedance measurements.....	6
1.4.2 APBRS and Parametric Chirp excitation, as amplitude modulation techniques for nonlinearity mitigation measure in BIS systems for solar modules	6
1.4.3 QFT signal injection control for the DC-DC converter to ensure accurate real-time measurements of a solar module under variable-point conditions.	7
1.4.4 Demonstration of a switch-mode converter’s inductor current ripple as an excitation signal to help reduce the switching frequency requirements of the SMC.....	8
1.5 Objectives and Limitations of the study.....	8
1.5.1 Solar module characterization.....	8
1.5.2 Wide bandwidth DC-DC boost SMC excitation source	8
1.5.3 Demonstration of amplitude modulation for device excitation signals	9
1.5.4 QFT signal injection control for the DC-DC converter to ensure accurate real-time measurements of a solar module under variable-point conditions.	9
1.5.5 Feasibility assessment into the use of the SMCs inductor current ripple as an excitation signal. 10	
1.5.6 Scope and limitations of the study	10
1.6 Research Outputs/Publications	11
1.7 Plan of development.....	12
1.7.1 Chapter 2.....	12
1.7.2 Chapter 3.....	12
1.7.3 Chapter 4.....	12
1.7.4 Chapter 5.....	13
1.7.5 Chapter 6.....	13

1.7.6	Chapter 7.....	13
1.7.7	Chapter 8.....	13
1.7.8	Chapter 9.....	13
1.7.9	Chapter 10.....	13
1.7.10	Appendices.....	13
1.8	References.....	14
Chapter 2	Photovoltaic module as the device under test	17
2.1	Introduction	17
2.2	Photovoltaic module electrical characteristics – an overview.....	18
2.2.1	Current and voltage characteristics.....	19
2.2.2	Module internal losses	20
2.3	Photovoltaic module direct current electrical equivalent circuit (DC EEC) models.....	21
2.3.1	Single diode model.....	21
2.3.2	Double diode solar cell model.....	22
2.3.3	Reverse bias model.....	23
2.3.4	Reverse bias model with dynamic effects.....	23
2.4	Photovoltaic module alternating current electrical equivalent circuit (AC EEC) models.....	25
2.4.1	First order Randel’s circuit model.....	25
2.4.2	Second order Randle’s circuit model	26
2.5	Photovoltaic module faults and degradation review	27
2.5.1	Partial shading.....	27
2.5.2	Micro cracks	27
2.5.3	Hot spots.....	28
2.6	Photovoltaic module diagnostics tests – a literature review.....	28
2.6.1	I/V curve testing.....	28
2.6.2	Radiometric sensors/thermography testing.....	29
2.6.3	Electroluminescence testing.....	29
2.6.4	Linear Impedance Spectroscopy testing.....	30
2.7	Chapter summary	31
2.8	References.....	32
Chapter 3	Linear impedance spectroscopy – a literature review	37
3.1	Introduction	37
3.2	Technical process high-level overview.....	37
3.3	Impedance representation and curve fitting.....	38
3.4	Device excitation signals.....	40
3.4.1	Introduction.....	40
3.4.2	Single frequency excitation signals.....	41

3.4.3	Broadband excitation signals	44
3.5	Excitation sources.....	46
3.5.1	AC current generator	46
3.5.2	Programmable electronics loads	47
3.5.3	Switch-mode converters	48
3.6	Process LTI conditions.....	50
3.6.1	Linearity	50
3.6.2	Causality	52
3.6.3	Time invariance	53
3.7	Impedance spectroscopy – an application review	53
3.7.1	Biomedical applications	53
3.7.2	Battery applications.....	54
3.7.3	Fuel cell applications	54
3.7.4	Solar module applications	55
3.8	Chapter summary	55
3.9	References.....	56
Chapter 4	Impedance extraction system design, analysis, and simulation results	62
4.1	Introduction	62
4.2	Impedance extraction system overview.....	63
4.3	Two-level boost SMC operational analysis.....	66
4.3.1	Single phase boost SMC	66
4.3.2	Multiphase boost SMC analysis	69
4.4	Excitation source boundary conditions for linearity	70
4.5	Two-level boost SMC power transfer analysis	72
4.6	Two-level boost SMC ripple analysis	74
4.6.1	Input inductor current ripple.....	74
4.6.2	Output capacitor voltage ripple.....	76
4.7	Two-level boost SMC active switch utilization analysis	77
4.7.1	Wide bandgap semiconductors – a high-level review	77
4.7.2	Voltage stress.....	78
4.7.3	Current stress	78
4.7.4	Switch utilization.....	79
4.8	Boost SMC power regulation and perturbation signal preliminary design.....	79
4.8.1	Power regulation design specifications	79
4.8.2	Solar module selection.....	80
4.8.3	Boost SMC power regulation design	81
4.9	Power regulation control – 1 st tier control design	83

4.9.1	Design specifications	84
4.9.1	System modelling	85
4.9.2	Closed loop system analysis.....	89
4.9.3	Controller design	92
4.10	Maximum power point tracking control design – 2 nd tier control design	97
4.11	Impedance spectroscopy duty cycle perturbation signals' design.....	98
4.11.1	Excitation signal design specifications.....	99
4.11.2	Chirp perturbation signal design.....	99
4.11.3	Triangular wave excitation signal design.....	100
4.11.4	PRBS perturbation signal design	100
4.12	Simulation results and discussions	102
4.12.1	Output voltage	102
4.12.2	Output voltage tracking specification.....	103
4.12.3	Maximum power point tracking.....	104
4.12.4	Load step.....	105
4.12.5	Input voltage step.....	106
4.12.6	Input current and output voltage ripple.....	107
4.12.7	Impedance spectroscopy voltage excitation signal induction.....	108
4.13	Chapter conclusions.....	113
4.14	References.....	114
Chapter 5 Experimental considerations and impedance extraction system design for online implementation.....		117
5.1	System setup overview.....	117
5.2	Frequency Response Analyser	119
5.3	Power optimizer multiphase boost converter	120
5.3.1	Prototype assembly design considerations.....	120
5.3.2	Power MOSFET Driver	122
5.3.3	Wide bandgap SiC semiconductor module configuration for high-speed performance	123
5.3.4	Active soft switching cell design	124
5.4	Power optimizer data acquisition and generation system.....	125
5.4.1	Transducers.....	126
5.4.2	Analog filters	126
5.4.3	Digital signal processing unit	128
5.4.4	DC component rejection	129
5.4.5	System's linearity during impedance measurements	129
5.4.1	Quantization error analysis	131
5.5	The embedded control unit algorithm design	133

5.6	Experimental results and discussions.....	135
5.6.1	Solar simulating light board.....	135
5.6.2	Solar module I/V curve characteristics.....	136
5.6.3	DC-DC power stage closed-loop results.....	137
5.6.4	Maximum power point tracking.....	143
5.6.5	Module optimal excitation signal design and bandwidth characterization.....	145
5.6.6	Impedance spectroscopy experimental voltage excitation signal induction.....	145
5.6.7	Offline Solar module impedance estimation results performed using a frequency response analyser	150
5.7	Chapter conclusions.....	158
5.8	References.....	160
Chapter 6 Rapid online solar module impedance estimation through SMC-based robust variable-point control.....		161
6.1	Introduction	161
6.2	System setup	162
6.3	Excitation parametric uncertainty analysis.....	163
6.4	Quantitative feedback control analysis.....	166
6.4.1	Output disturbance rejection	167
6.4.2	Input disturbance rejection	167
6.4.3	Noise rejection.....	168
6.4.4	Tracking specification	168
6.4.5	Robust stability	168
6.5	Quantitative feedback control synthesis.....	169
6.5.1	Design specifications	169
6.5.2	Choice of frequency array.....	170
6.5.3	Choice of nominal plant	170
6.5.4	Plant template generation	170
6.5.5	Full system design.....	170
6.6	Experimental results and discussions.....	174
6.6.1	Induced voltage excitation signal results during variable-point conditions.....	174
6.6.2	Impedance measurement results under variable conditions.....	175
6.7	Chapter conclusions.....	178
6.8	References.....	178
Chapter 7 Rapid online module impedance estimation through SMC-based parametric chirp excitation 180		
7.1	Introduction	180
7.2	System setup.....	181

7.3	Chirp excitation analysis and design	182
7.3.1	Chirp signal time-domain analysis.....	182
7.3.2	Linear chirp.....	182
7.3.3	Geometric chirp.....	183
7.3.4	Chirp signal frequency-domain analysis	184
7.4	Amplitude modulated parametric chirp excitation analysis and design	185
7.5	Experimental results and discussions.....	186
7.5.1	Chirp signal frequency domain results	187
7.5.2	Parametric chirp excitation signal induction analysis.....	187
7.5.3	Impedance estimation results.....	188
7.6	Chapter conclusions.....	189
7.7	References.....	190
Chapter 8 Rapid online solar module impedance estimation through SMC-based amplitude-modulated pseudo-random binary sequence excitation		192
8.1	Introduction	192
8.2	APRBS excitation system setup.....	193
8.3	PRBS amplitude modulation design.....	193
8.4	Experimental results and discussions.....	198
8.4.1	Voltage excitation signal induction.....	198
8.4.2	Impedance results.....	199
8.5	Chapter conclusions.....	202
8.6	References.....	202
Chapter 9 Solar module impedance estimation through SMC-based inductor current ripple excitation 204		
9.1	Introduction	204
9.2	Excitation system overview.....	205
9.3	Boundary between external and current ripple excitation	206
9.3.1	Choice of converter switching speed	206
9.3.2	Characterization under high irradiance conditions	207
9.3.3	Characterization under low irradiance conditions	208
9.4	Design considerations for the current ripple excitation magnitude	209
9.5	Design considerations for the excitation frequency.....	211
9.6	Performance assessment during foldback control	212
9.6.1	System efficiency	212
9.6.2	Input current ripple.....	212
9.6.3	Output voltage ripple.....	213
9.6.4	Load transient response.....	214

9.6.5	Loop stability	214
9.7	Optimal frequency points	215
9.7.1	Series resistance	215
9.7.2	Diffusion resistance	216
9.7.3	Diffusion capacitance	217
9.8	System design for inductor current ripple excitation	217
9.9	Experimental results and discussions	218
9.9.1	Current ripple variation and foldback control results	218
9.9.2	Impedance results	219
9.10	Chapter conclusions	222
9.11	References	223
Chapter 10	Conclusions and recommendations	225
10.1	Concluding remarks	225
10.1.1	Satisfactory development of a wide bandwidth multiphase converter	225
10.1.2	Satisfactory implementation of perturbation signal amplitude modulation	226
10.1.3	Satisfactory impedance extraction control system	226
10.1.4	Satisfactory evaluation of crystalline solar modules' impedance as the DUT	227
10.1.5	Satisfactory performance for linear impedance spectroscopy using an SMC on a solar PV module as a characterization technique	228
10.1.6	Relevance of impedance extraction system to industry applications	228
10.2	Study recommendations	229
10.2.1	Assess other module types	229
10.2.2	Assess other conditions and faults	229
10.2.3	Investigate alternate processing units.	229
10.3	Future work recommendations	229
10.3.1	Simultaneous Impedance estimation of solar module and battery storage	230
10.3.2	Partial shading assessment through the impedance characterization of the bypass diodes	230
10.3.3	Investigate the use of a square wave frequency sweep signal for reduced signal processing power requirements	231
10.3.4	Investigate the use of a multiphase converter to control inductor current ripple magnitude during device excitation	231
10.3.5	High frequency impedance prediction algorithm	232
Appendices		233

List of Figures

Figure 2.1 Generalised solar module current and voltage characteristics.....	17
Figure 2.2 Single diode static model for a photovoltaic cell.....	20
Figure 2.3 Double diode static model for a polycrystalline cell.....	20
Figure 2.4 Static solar cell reverse bias model considering the avalanche effect.....	21
Figure 2.5 PV cell dynamic reverse bias model	22
Figure 2.6 First order Randel's AC-EEC model for a photovoltaic cell.....	23
Figure 2.7 Second order Randel's AC-EEC model for a photovoltaic cell.....	24
Figure 2.8 Residential solar PV system condition monitoring using DC-DC power optimizers.....	26
Figure 3.1 Linear impedance spectroscopy high-level process overview	36
Figure 3.2 Electrical equivalent circuit modelling from impedance spectroscopy data (a) series RL (b) series RC (c) parallel RC (d) series R-RC	37
Figure 3.3 Device excitation signal for system identification review	38
Figure 3.4 Sinusoidal excitation signal ideal frequency response	39
Figure 3.5 Triangular wave excitation signal ideal frequency response (a) 100% symmetry (b) 0% symmetry.....	41
Figure 3.6 Twelve-bit PRBS sequence generator.....	43
Figure 3.7 Frequency spectrum of PRBS sequence.....	43
Figure 3.8 AC current generator device excitation	44
Figure 3.9 Electronic load device excitation	45
Figure 3.10 Converter device excitation.....	46
Figure 3.11 Block diagram illustrating linearity of an electronic system.....	48
Figure 3.12 Illustration of pseudolinearity from an arbitrary electrochemical device	48
Figure 3.13 Optimal perturbation amplitude for a monocrystalline solar module as a percentage of its DC voltage.....	49
Figure 3.14 Block diagram illustrating causality of an electronic system	50
Figure 3.15 Block diagram illustrating time invariance of an electronic system	50
Figure 4.1 Proposed system block diagram demonstrated as a power optimizer for online impedance extraction of a solar module (a) Location of the proposed system on a residential PV system (b) Detailed module-level system description.....	62
Figure 4.2 Boost switch-mode DC-DC converter topology	64
Figure 4.3 Boost SMC CICM modes of operation (a) ON-state (b) OFF-state	65
Figure 4.4 Single-phase boost SMC switch, inductor, and capacitor current signals at 50% duty cycle using PWM	65
Figure 4.5 Multiphase boost switch-mode DC-DC converter topology.....	66
Figure 4.6 Multiphase boost SMC operational analysis (a) Mode 1 (b) Mode 2	67

Figure 4.7 Two-phase multiphase boost SMC switches, inductors, and capacitor current signals at 50% duty cycle using PWM	67
Figure 4.8 Boost SMC voltage transfer ratio (a) varying inductor parasitic resistance (b) varying load current	71
Figure 4.9 Peak to peak input current ripple comparison between the single and double phase	73
Figure 4.10 Peak to peak output voltage ripple comparison between the single and double phase	74
Figure 4.11 Shell PowerMax I/V characteristics	78
Figure 4.12 Shell PowerMax P/V characteristics	78
Figure 4.13 Control system closed loop architecture	87
Figure 4.14 Voltage plant closed loop response with unity gain (a) Bode response (b) root locus (c) step response	87
Figure 4.15 Current plant closed loop response with unity gain (a) Bode response (b) root locus (c) step response	88
Figure 4.16 Lead compensated current loop response (a) Bode response (b) root locus (c) step response	92
Figure 4.17 Stages in the induction of the excitation and response signals	93
Figure 4.18 Solar module AC EEC frequency response under several operation conditions (a) gain (b) phase	93
Figure 4.19 PI compensated voltage loop response (a) Bode response (b) root locus (c) step response	94
Figure 4.20 Incremental conductance maximum power point tracking algorithm flowchart for the module	95
Figure 4.21 Steady state boost converter input and output voltage (a) single phase (b) multiphase ..	100
Figure 4.22 Charging voltage tracking specification	101
Figure 4.23 System's maximum power point tracking with varying irradiance	102
Figure 4.24 System's maximum power point tracking with varying temperature	102
Figure 4.25 Load disturbance rejection (a) output voltage (b) output current	103
Figure 4.26 System input voltage disturbance test (a) output voltage (b) input voltage	104
Figure 4.27 Boost converter input current ripple (a) single phase (b) multiphase	105
Figure 4.28 Boost converter output voltage ripple	105
Figure 4.29 Chirp excitation analysis	107
Figure 4.30 Chirp excitation frequency comparative analysis (a) perturbation duty cycle signal (b) voltage excitation signal	107
Figure 4.31 Triangular wave inductor current excitation and voltage response signals at varying frequency (a) 20kHz (b) 40kHz (c) 70kHz (d) 100kHz	108
Figure 4.32 PRBS modulating duty cycle perturbation signal (a) PRBS time domain signal (b) perturbation signal frequency power spectrum	109

Figure 4.33 Induced voltage signal results (a) excitation time signal (b) excitation signal frequency power spectrum.....	110
Figure 5.1 Online impedance extraction system laboratory setup.....	115
Figure 5.2 Online impedance extraction system setup block diagram	115
Figure 5.3 AUTOLAB PGSTAT302N frequency response analyser block diagram	116
Figure 5.4 SiC MOSFET high-frequency evaluation board block diagram.....	119
Figure 5.5 Silicon Carbide MOSFET module configuration for high performance bipolar switching with Miller clamp.....	120
Figure 5.6 Boost converter per-phase active soft switching cell.....	120
Figure 5.7 Data acquisition system setup block diagram	121
Figure 5.8 Sallen-key second order low pass filter configuration	123
Figure 5.9 DC component rejection control technique.....	125
Figure 5.10 Embedded control unit flow algorithm	130
Figure 5.11 Solar simulating light board performance with perpendicular distance (a) irradiance (b) temperature.....	132
Figure 5.12 Module I/V curve characteristics of the monocrystalline module (a) varying irradiance (b) varying temperature.....	133
Figure 5.13 Steady state boost converter input and output voltage (a) single phase (b) multiphase..	135
Figure 5.14 Output voltage performance under voltage setpoint commands.....	136
Figure 5.15 Output voltage performance under load disturbance	137
Figure 5.16 Boost converter input current ripple (a) single phase (b) multiphase	138
Figure 5.17 Comparative efficiency analysis between single-phase and multiphase converter with varying load power.....	140
Figure 5.18 Module incremental conductance power point tracking under varying irradiance.....	140
Figure 5.19 Response current THD resulting from voltage perturbations of various amplitudes, as a percentage of the bias voltage	141
Figure 5.20 Induced chirp voltage excitation signal under input voltage variation	144
Figure 5.21 Induced chirp voltage excitation signal under load variation.....	144
Figure 5.22 Induced PRBS voltage excitation signal under input voltage variation	145
Figure 5.23 Induced PRBS voltage excitation signal under load variation	145
Figure 5.24 Solar module inductor current excitation (a) 20kHz (b) 100kHz	146
Figure 5.25 Points on the I/V curve at which varying operating point impedance measurements were taken	147
Figure 5.26 Module Nyquist impedance response under high irradiance conditions and varying temperature.....	148
Figure 5.27 Points on the I/V curve at which varying irradiance impedance measurements were taken	149

Figure 5.28 Module Nyquist impedance response under varying irradiance levels.....	150
Figure 5.29 Points on the I/V curve at which varying temperature impedance measurements were taken	151
Figure 5.30 Nyquist plot showing temperature effects on the impedance of a solar module.....	152
Figure 5.31 Points on the I/V curve at which partial shading impedance measurements were taken	153
Figure 5.32 Module Nyquist impedance response under partial shading conditions.....	153
Figure 5.33 Module Nyquist impedance response under microcracks	154
Figure 6.1 Quantitative feedback control excitation system setup	160
Figure 6.2 Effect of system uncertainties on the induced excitation broadband signal (a) Condition dependent parameters (b) BIS frequency	162
Figure 6.3 Inductance variation with average current and temperature at the switching frequency .	163
Figure 6.4 QFT control system 2-degree of freedom closed loop for voltage excitation signal induction	163
Figure 6.5 QFT synthesis (a) Templates (b) Time domain tracking specification (c) Frequency domain tracking specification (d) Tracking bounds (e) Stability and sensitivity bounds (f) Intersectional bounds with control law achieved through loop shaping	170
Figure 6.6 Induced voltage excitation signals during various operating points under quantitative feedback excitation, showing input and noise disturbance rejection, tracking, and robust stability ...	172
Figure 6.7 Induced voltage excitation signals during load changes under quantitative feedback excitation, showing output and noise rejection and robust stability	172
Figure 6.8 Impedance estimation on a Nyquist plot under varying conditions without QFT control injection (a) irradiance variation (b) temperature variation	173
Figure 6.9 Impedance estimation shown on a Nyquist plot under varying conditions with QFT control injection (a) irradiance variation (b) temperature variation	173
Figure 6.10 Variation of module parameters with irradiance (a) resistive (b) capacitive.....	174
Figure 6.11 Variation of module parameters with temperature (a) resistive (b) capacitive.....	174
Figure 7.1 Amplitude-modulated chirp excitation system setup	180
Figure 7.2 Spectrogram analysis for chirp excitation signals (a) linear and (b) geometric.....	185
Figure 7.3 Parametric module chirp excitation.....	186
Figure 7.4 PV system impedance estimation under normal well-lit conditions (a) Nyquist plot (b) Bode plot	187
Figure 7.5 PV system impedance estimation under hotspots (a) Nyquist plot (b) Bode plot.....	187
Figure 7.6 PV system impedance estimation under partial shading conditions (a) Nyquist plot (b) Bode plot	187
Figure 8.1 Amplitude-modulated pseudorandom binary sequence parametric excitation system setup	192

Figure 8.2 APRBS signal generation (a) 12-bit APRBS signal (b) APRBS slope (c) APRBS power spectrum	196
Figure 8.3 Signal input space distribution (a) PRBS (b) APRBS	197
Figure 8.4 Experimental induced voltage excitation signals from modulating duty cycle perturbations (a) PRBS (b) APRBS.....	198
Figure 8.5 Induced APRBS voltage excitation signal resulting from APRBS perturbation, under input voltage variation.....	198
Figure 8.6 Nyquist APRBS impedance extraction of a PV module under microcracks and well-lit conditions (a) high irradiance (b) low light.....	199
Figure 8.7 APRBS impedance magnitude of a PV module under microcracks and low-lit conditions (a) high irradiance (b) low irradiance	199
Figure 8.8 Impedance estimation during partial shading (a) Nyquist plot (b) Impedance magnitude	200
Figure 9.1 Inductor current ripple excitation system setup	205
Figure 9.2 1D frequency distribution of the module's Nyquist response under high irradiance conditions	207
Figure 9.3 Frequency distribution of the module's Nyquist response under low irradiance conditions	208
Figure 9.4 Durban direct irradiance profile during the winter and summer solstices [5]	209
Figure 9.5 Effects of the inductor's parasitic resistance on the magnitude of the peak-peak inductor current ripple.....	210
Figure 9.6 Graphical analysis of a typical Nyquist impedance representation for a first order electrochemical device.....	215
Figure 9.7 Boost converter input inductor current ripple at various switching speeds	218
Figure 9.8 Foldback control of a boost converter's inductor current ripple	219
Figure 9.9 Series resistance under variations in (a) irradiation - well lit (red) low light (blue) (b) temperature - low (red) high (blue).....	220
Figure 9.10 Inductor current ripple impedance estimation under high irradiance conditions	220
Figure 9.11 Inductor current ripple impedance estimation under low irradiance conditions	221
Figure 10.1 AUTOLABPGSTAT302N frequency response analyser	233
Figure 10.2 Multiphase boost power optimizer prototype top view.....	234
Figure 10.3 Multiphase boost power optimizer prototype side view	234
Figure 10.4 CRD-5FF0912P, SiC MOSFET High-Frequency Evaluation Board for 7L D2PAK top view.	235
Figure 10.5 CRD-5FF0912P, SiC MOSFET High-Frequency Evaluation Board for 7L D2PAK side view.	235
Figure 10.6 Single phase boost converter model showing impedance estimation.....	236
Figure 10.7 Multiphase phase boost converter model showing impedance estimation.	236

Figure 10.7 Power regulation control MATLAB graphical model.....	236
Figure 10.7 MPPT control MATLAB graphical model.	237

List of tables

Table 2.1 Comparison of different solar cell types.....	16
Table 4.1 Comparison of conventional and wide bandgap semiconductor properties	76
Table 4.2 Residential hybrid solar system with storage specifications.....	77
Table 4.3 Control design specifications	81
Table 4.4 Induced voltage excitation signal design specifications	96
Table 5.1 Solar module test power matrix.....	133
Table 5.2 Output voltage performance comparison under several voltage setpoint commands.....	136
Table 5.3 Output voltage performance comparison under load disturbance test	137
Table 5.4 Efficiency analysis of a single-phase boost converter.....	139
Table 5.5 Efficiency analysis of a multiphase boost converter	139
Table 5.6 Monocrystalline solar module parameter estimation under variable operating points	148
Table 5.7 Monocrystalline solar module parameter estimation under varying irradiance	150
Table 5.8 Monocrystalline solar module parameter estimation under temperature changes	152
Table 6.1 QFT design specifications	166
Table 6.2 Nominal plant model and its parametric uncertainty changes	166
Table 8.1 Error estimation under microcracks in high irradiance conditions	200
Table 8.2 Error estimation under microcracks in low irradiance conditions.....	200
Table 8.3 Error estimation under partial shading.....	201
Table 9.1 Inductor current ripple excitation system power regulation specifications.....	217
Table 9.2 Monocrystalline solar module parameter estimation from the inductor current ripple.....	221

List of abbreviations

AC	Alternating Current
ADC	Analog to Digital Converter
AI	Artificial Intelligence
AM	Air Mass
APRBS	Amplitude-modulated Pseudorandom Binary Sequence
BESS	Battery Energy Storage System
BIS	Broadband Impedance Spectroscopy
BLDC	Brushless DC
BMS	Battery Management Systems
CICM	Continuous Inductor Current Mode
DAC	Digital to Analog Converter
DC	Direct Current
DCR	Direct Current Resistance
DFT	Discrete Fourier transform
DoE	Design of Experiments
DUT	Device Under Test
ECU	Embedded Control Unit
EEC	Electrical Equivalent Circuit
EIS	Electrical Impedance Spectroscopy
EMI	Electromagnetic Interference
EPC	Engineering, Procurement and Construction
ESR	Equivalent Series Resistance
FF	Fill Factor
FFT	Fast Fourier Transform
FIR	Finite Impulse Response
FMH	Finite Median Hybrid
FPGA	Field-Programmable Gate Array
FRA	Frequency Response Analyser
GaN	Gallium Nitride
GM	Gain Margin
LIS	Linear Impedance Spectroscopy
LTI	Linear Time-Invariant
MIMO	Multiple Input Multiple Output
MPPT	Maximum Power Point tracker

NI	National Instruments
NLEIS	Nonlinear Electrical Impedance Spectroscopy
PE	Power Electronic
PEL	Programmable Electronic Load
PEM	Proton Exchange Membrane
PFM	Pulse Frequency Modulation
PID	Potential Induced Degradation
PM	Phase Margin
PoC	Point of Connection
PPC	Power Plant Controller
PRBS	Pseudorandom Binary Sequence
PWM	Pulse Width Modulation
QFT	Quantitative Feedback Theory
RHP	Right Half Plane
SCADA	Supervisory Control and Data Acquisition
SFRA	Sweep Frequency Response Analysis
Si	Silicon
SiC	Silicon Carbide
MIMO	Multiple Input Multiple Output
SMC	Switch-Mode Converter
SNR	Signal to Noise Ratio
SoC	State of Charge
SoH	State of Health
SSEG	Small Scale Embedded Generation
STFT	Short-Time Fourier Transform
THD	Total Harmonic Distortion
WBG	Wide bandgap

Chapter 1

Introduction

1.1 Background to the study

The rise in global carbon emissions has prompted research to reduce the earth's carbon footprint. One of the areas of interest is utilising sustainable energy resources such as wind and solar energy – instead of burning fossil fuels. The renewable energy market has grown exponentially over the past couple of years because advances in technology has seen the prices for these systems decrease. One of the ways of harvesting renewable energy, is by making use of solar Photovoltaic (PV) modules. In 2020, renewable energy accounted for 29% of global energy production, a 13% increase from the previous year; in renewable energy production, solar energy accounts for 32% of this production [1] – [3]. PV installations reached more than 821Wh in 2020, a growth of more than 23% from 2019 [4]. Even though the cost of PV systems is decreasing annually, these systems remain expensive for average consumers. One of the reasons for this large cost is a large operation and maintenance cost associated with small scale PV systems; the operation and maintenance costs are among the highest, accounting for more than 20% of solar system costs; this bracket includes, hardware and software installations, testing and monitoring the module performance [5]; because of this, an advancement in existing technologies is required to improve condition monitoring of Small-Scale Embedded Generation (SSEG) solar PV systems.

1.2 Problem statement

Solar PV systems have traditionally relied on terminal DC current and voltage characteristics for online diagnostics. This approach monitors the current and voltage levels of a module, and a change in these levels indicates a system anomaly. However, this information only alerts to a decrease in the current or voltage levels, but not the possible cause(s) for the anomaly – whether the module is partially shaded; the cells have degraded; the cells have cracked; or an increase in wiring resistance due to hot

spots, among other faults. This information is vital since most PV modules are mounted at high elevations and are accessible only through physical interaction, which contributes to increased operation and maintenance costs.

Detailed online tests can be taken using thermography-based methods that use unmanned aerial vehicles to obtain more information while solar modules are operating [7]; they are, however, limited to heat-sensitive degradation, such as hot spots, large cell cracks, and partial shading, but do not offer continuous online diagnostics. Electroluminescence is another technique employed which provides detailed diagnostics – particularly for microcracks and hot spots, however, the technique suffers from a lack of online diagnostics because solar modules must be taken offline [8]. Therefore, to improve online diagnostics, what is required is an online diagnostics system that can provide detailed information – regarding the drop in PV module performance – without requiring special equipment, like thermography and electroluminescence-based techniques; this will improve online diagnostics of a solar module without increasing monitoring system costs and help improve online diagnostics and pre-emptive maintenance of a solar module.

Research has shown how frequency-based techniques, such as Linear Impedance Spectroscopy (LIS) – a primarily lab-based technique implemented through Frequency Response Analyzers (FRAs) for material selection and optimization – can be implemented on electrochemical devices to monitor the state of health through impedance characterization [9] – [11]. The technique uses single frequency sinusoidal perturbations to characterize a device's impedance – also referred to as Electrical Impedance Spectroscopy (EIS). The results presented in the research showed that LIS provides good diagnostics, with a significant reduction in cost and a reduction in data acquisition time to minutes, compared to other lab-based techniques – which can take several hours or days in some cases – thus improving the reliability of diagnostics for these devices [9] – [11].

Because of the single frequencies used in EIS however, real-time/online measurements cannot be conducted linearly using the technique because the data acquisition time can go up to 100s in cases where more frequency points are required; in that much time, the device's internal operating point can change significantly, thus linearity would not be observed in the measurements. Thus, to further reduce data acquisition time and implement LIS in real-time/online, research has shown how broadband frequency perturbations i.e., Broadband Impedance Spectroscopy (BIS) – implemented through DC-DC power electronic (PE) switch-mode converters (SMC) – can reduce the data acquisition time for the measurements to as short as 3s in batteries and fuel cells[12] – [15], thus allowing for feasible real-time/online diagnostics. BIS differs from EIS in that it injects multiple perturbation frequencies at an instance – this drastically decreases the data acquisition time. With LIS implemented as BIS through SMCs, it can be implemented outside laboratory settings and in the field where batteries and fuel cells are operational. The results from the research showed that the BIS online diagnostics system can help drastically reduce diagnostics acquisition time, costs, system downtimes, and offer pre-emptive correction to failures that are not generally detectable at their infancy [16] – [18].

EIS in a lab setting has been implemented on solar PV modules and research showed that this impedance information can then be used to determine underlying faults such as partial shading, hot spots, cracks, and PID, cell degradation, ageing, mechanical stress, which combines a wider range of faults compared to the other techniques for PV [19] – [23]. Because of the slow nature of EIS, recent research advancements in implementing LIS on PV modules include the use of BIS through PE loads to reduce the data acquisition time and enable online impedance measurements; however, this advancement is still only applicable to lab-based measurements because of the use of PE loads [24] – [26]. Thus, although the online implementation of BIS through a DC-DC SMC has been researched and thoroughly demonstrated on batteries and fuel cells, current research has not thoroughly researched or demonstrated the technique on a solar module; this is because on a solar module, the implementation presents challenges that have not yet been identified or addressed in research; these challenges are discussed in this section and will serve as the core focus of the study.

1.2.1 *The high impedance bandwidth of solar PV modules*

Converters used in power regulation typically rely on traditional silicon switches, which can operate at low switching frequencies like $20kHz$. This is usually sufficient in impedance estimation systems for batteries and fuel cells due to their low impedance bandwidths, which are less than $6kHz$ for batteries [16][17] and less than $10kHz$ for fuel cells [19]. The impedance bandwidth of a device refers to all frequency responses that result in a total harmonic distortion 8% in a current or voltage response signal from an injected perturbation voltage or current signal [18]. However, crystalline solar modules, have a higher bandwidth than batteries, greater than $10kHz$ and up to $100kHz$ [19] – [26]; and this presents a challenge in extracting useful information from PV modules. Thus, to accurately reconstruct the BIS excitation signal using the switching signal of the SMC, the frequency of the switching signal must be at least 10 times higher than the bandwidth of the PV module. This has several implications on the impedance measurements of a solar module, the converter's design, switching, and control designs, that have not been addressed.

1.2.2 *Linearity of the excitation signals under induced excitation in high bandwidth systems*

The use of broadband signals for device characterization through a switch-mode converter can only be achieved through induced excitation [16],[17],[20], [24] – [26]; that is, the device excitation is induced from an externally generated perturbation duty cycle signal. This requires an in-depth analysis into the input-output relationship between the duty cycle perturbation broadband signal and the induced broadband voltage excitation signal – most importantly, how the frequency response of this relationship affects the characteristics of the induced broadband excitation signal. Current research into converter-based impedance measurements, focuses on optimizing the amplitude of the excitation signal for device linearization without factoring in induced excitation – a factor that is unique when using an SMC for impedance measurements and needs to be addressed to improve the accuracy of online diagnostics using BIS.

To further decrease measurement time and reduce the system processing power requirement, binary broadband signals have been investigated on battery systems in [21] – [23]. The results have shown even faster measurements times and a reduction in systems costs. The problem with using these binary signals, however, is that both the converter and electrochemical device have non-linear behavior; this non-linearity causes larger impedance errors at certain frequencies and hence, a major issue for wide bandwidth systems like PV systems. Another factor current research, into binary excitation for converter-based impedance spectroscopy has not addressed.

The inductor current ripple of the SMC used as an excitation signal, for a solar PV system, is another area of research that has not been addressed; this is because of the complexity that comes with designing a signal that can be responsible for impedance measurements without impacting power regulation. The peak-peak inductor current ripple magnitude must be designed to ensure linearity in impedance measurements without increasing the capacitor and inductor components sizes of the SMC; the switching frequency of converter must be varied to measure impedances at different frequencies with minimal impact to power regulation. With these 2 challenges solved, the inductor current ripple as an excitation signal will help solve the issue of requiring a high switching frequency in the SMC.

1.2.3 Control of online impedance extraction under variable-point conditions

Although limited, there have been a few effective control techniques that have been implemented in online impedance estimation systems. In [16], [17] and [20] – [22], hysteresis-type control algorithms were implemented in online impedance estimation for battery systems; the control systems were designed based on certain steady state conditions of the output voltage before impedance measurements were taken. This type of control works because if the signal is injected while the system is not at steady state, the impedance measurements of the devices would not be taken at a single operating point. The flaw of this type of control, however, is that it is designed for a single operating point of measurement such that the injected signal is left uncompensated for any system variations. Current control in [27] and resonant control in [28] were implemented, both these techniques also suffer from single point control. Thus, the common issue with existing control techniques employed for impedance measurements is summarized as follows.

When controlling the excitation signals through the converter in real-time, the excitation frequency, solar module's impedance, and switch-mode converter's parameters form part of the dynamic system describing the small signal plant, from the injected duty cycle modulating signal to the induced voltage excitation signal at the input of the converter. Since the solar module and converter's parameters vary with changing weather conditions, the dynamic system also changes; when this happens, the excitation signal will change; existing control systems for online impedance measurements have not considered this fact. This is a major issue because the excitation signal must remain within a certain linear range during the measurements. If not properly controlled, this can lead to large excitation signal amplitudes that push the system to a non-linear range or result in amplitudes that are too small for impedance estimation. These are issues that existing control techniques employed for impedance extraction cannot

address because of the type of architecture they use. What is required, is a control technique that can allow for a wide variation of operating points within the conditions under which a typical solar module operates during signal injection.

1.3 Research questions

Based on the problems identified, the research questions considered were as follows:

- i. *What is the relation between the solar module's operational or fault condition and its AC impedance?*
- ii. *What DC-DC SMC configuration is most suitable to operate as an excitation source for solar module impedance extraction?*
- iii. *What is the maximum achievable loop bandwidth by a boost SMC excitation source?*
- iv. *How can the amplitude profile of an analog excitation signal be designed to limit the nonlinearity effect of induced excitation?*
- v. *How can a digital excitation signal be designed for solar PV module excitation while observing linearity in the measurements?*
- vi. *To what degree can the inductor current ripple of the SMC be used for impedance measurements of a solar PV module?*
- vii. *How can impedance measurements be taken accurately when the operating point of the SMC varies?*

1.4 Purpose and contributions of this study

This study thus, presents an advancement to converter-based online AC impedance extraction system for PV modules through broadband impedance spectroscopy – impedance information that can be used for online diagnostics of a solar module. This system can be integrated in either a hybrid power optimizer or hybrid charge controller to perform module-level online impedance measurements on a module, while it optimizes power delivered to a load. With the proposed advancement, these devices can be designed to extract a solar module's AC impedance, such that any change in this impedance will reveal more information about the module's performance changes. In real-world PV systems, this AC impedance can help improve online diagnostics of a solar module by offering pre-emptive maintenance and a lower cost solution to detailed online diagnostics.

Based on the problems identified, to develop the impedance extraction system for a solar PV module through a DC-DC SMC, 3 major components are required: sufficient bandwidth in the SMC power regulator to accommodate a solar PV module's wide impedance bandwidth while efficiently regulating power; excitation signals with frequency induced nonlinearity mitigation to ensure linearity in the impedance measurements; and a signal injection control system which produces induced excitation signals

that can keep the solar module in its linear range of operation while the operation point of the system varies. Thus, as part of 5 major contributions, the study will propose.

- i. *A wide-bandwidth multiphase boost excitation source for a solar module to perform high-frequency impedance measurements.*
- ii. *Parametric Chirp excitation, as an amplitude modulation technique for nonlinearity mitigation measure in converter-based BIS impedance extraction of a solar module.*
- iii. *Amplitude-modulated pseudorandom binary sequence excitation, as a solution to linearly implement digital excitation in converter-based BIS impedance extraction of a solar module.*
- iv. *A novel excitation signal injection control based on QFT control, for impedance measurements under variable-point conditions, which is part of 3 tier control system to execute converter-based BIS impedance extraction of a solar module.*
- v. *Demonstration of a switch-mode converter's inductor current ripple as an excitation signal to help reduce the switching frequency requirements of the SMC.*

These areas have not been addressed or investigated in research for online AC impedance estimation for solar modules. Thus, the details of these contributions are discussed in this section.

1.4.1 *A wide-bandwidth multiphase boost excitation source for a solar module to perform high-frequency impedance measurements.*

As a first contribution, this study will demonstrate the use of a wide bandwidth multiphase boost DC-DC SMC as an excitation source, to regulate solar power delivered to a load battery while performing impedance measurements on a solar module. A traditional multiphase boost converter will be modified for high-frequency performance using techniques discussed in the objectives; as well as using SiC switches to help produce a 100kHz loop bandwidth, and to ensure efficient power regulation while performing impedance measurements of a solar module. SiC semiconductors are wide bandgap devices that have been shown to provide better switching performance at higher frequencies [28] – [30]. The multiphase configuration was specifically chosen to reduce the high switching frequency input current ripple – an undesirable component in the application. Such modifications are specific to AC impedance extraction of a crystalline solar module through a DC-DC SMC, based on the research conducted in this thesis, and are thus, unique to the application.

1.4.2 *APBRS and Parametric Chirp excitation, as amplitude modulation techniques for nonlinearity mitigation measure in BIS systems for solar modules*

As part of 2 more major contributions, this study will demonstrate the use of perturbation signal amplitude modulations in both analog chirp and digital PRBS excitation signals, to mitigate frequency-induced nonlinearities in the SMC's induced voltage excitation signals, while performing broadband impedance measurements on the PV module.

A Parametric Chirp excitation signal will be proposed, which will be achieved through the frequency-domain convolution between the chirp perturbation signal and the power converter's dynamic small signal response – which includes the solar module's dynamic response. This will ensure that the amplitude profile of the perturbation signal can provide a mitigating response to high gains at low frequencies, and low gains at high frequencies.

An Amplitude-modulated Pseudorandom Binary Sequence (APRBS) excitation signal will be proposed and will take the form of discrete design points synthesized using an adaptive model-based online design of experiments (DoE). The input space for signal amplitude optimization will be defined by the non-linear relationship between the converter's PRBS modulating duty cycle and the converter's induced PRBS voltage excitation signal for adaptive nonlinearity mitigation. APRBS excitation will provide faster impedance estimation results, but with better frequency induced system non-linearity mitigation across a wide bandwidth and ensure that the signal can provide accurate impedance results over various system operating points.

1.4.3 QFT signal injection control for the DC-DC converter to ensure accurate real-time measurements of a solar module under variable-point conditions.

As the 4th major contribution, this study will demonstrate the synthesis of a novel excitation signal injection control, realized through quantitative feedback control theory (QFT). QFT is a variable-point control technique that has been shown to characterize and mitigate a system's structured uncertainties (uncertainties with known variations) on converters and other applications. The signal injection control will ensure that the induced voltage excitation signals have an amplitude range that keeps the system linear for the duration of the impedance measurements for better online impedance measurements on a solar module as its operating point varies.

The signal injection control will be implemented as part of a 3-tier control system, consisting of power regulation control, MPPT control, and QFT variable-point signal injection control, to ensure regulated power and impedance measurements are taken accurately when the PV, SMC, and load conditions vary. The power regulation control will take the form of a double-loop architecture comprising of a nested input current and output voltage loop for robust power transfer. The nested current loop will be responsible for ensuring that the inductor is properly controlled to ensure power stability; this loop will also be responsible for the excitation signal injection and will thus be designed to achieve a wide loop bandwidth through lead circuit compensation.

An incremental conductance MPPT algorithm will be designed to optimize the module's power, achieve good MPP convergence times, and minimal oscillations during maximum power to accurately inject the perturbation signals. The MPPT control will also be used to determine whether the module is operating as a current or voltage during impedance measurements, by comparing the operating current or voltage to the maximum current and voltage under those conditions; that is, the module will operate

as a current source for all currents larger than the current at maximum power – which is conversely true for the voltage.

1.4.4 Demonstration of a switch-mode converter's inductor current ripple as an excitation signal to help reduce the switching frequency requirements of the SMC.

As a 5th contribution, this study will undertake a feasibility analysis to assess the degree to which the inductor current ripple of a boost SMC can be used to perform impedance measurements on the solar module. This differs from traditional LIS implemented through an SMC because the excitation signal is not induced through an externally generated perturbation duty cycle signal but generated within the SMC's peak to peak inductor current ripple. With LIS implemented in this manner, the switching frequency of the converter will only be required to match the highest frequency in the DUT's impedance bandwidth – in the case of the solar module in this study, 100kHz.

1.5 Objectives and limitations of the study.

The objectives for each of the contributions discussed in the previous section, the objectives related to the solar module and its impedance measurements, as well as the limitations of the study, are discussed in this section.

1.5.1 Solar module characterization

- i. Determine the optimal LIS amplitude for the chosen solar PV module, using FRA, by monitoring the THD of the current response signals from injected voltage signals.*
- ii. Determine the impedance bandwidth of the solar module, by performing offline EIS on the solar module using the FRA.*
- iii. Use the FRA to create a benchmarking set of impedance curves to compare with the designed impedance extraction system.*

1.5.2 Wide bandwidth DC-DC boost SMC excitation source

These are the detailed objectives adopted to successfully design and build the wide bandwidth SMC for the extraction system.

- i. Analyse and design both single and multiphase boost SMCs to perform EIS and BIS impedance measurements while providing regulated power to a DC load.*
- ii. Design a **100kHz** current loop bandwidth power regulation control system.*
- iii. Assess the impact of voltage loop non-minimum phase system on impedance measurements.*

- iv. *Design and build a resonant soft switching cell to allow for efficient switching at 500kHz.*
- v. *Configure the gate drive circuit for high-performance bipolar switching at 500kHz to improve the switching utilization.*
- vi. *Design and build DC-DC SMC prototype and test its power regulation and signal injection capabilities for impedance extraction.*

1.5.3 Demonstration of amplitude modulation for device excitation signals

These are the detailed objectives adopted to design, analyze, and demonstrate the benefit to perturbation signal amplitude modulation.

- i. *Analyze and design for the optimal excitation frequency of a chirp signal.*
- ii. *Analyze and design for the optimal excitation frequency of a PRBS signal.*
- iii. *Model the power converter and solar module frequency responses.*
- iv. *Perform frequency-domain convolution for parametric chirp broadband excitation as a 2nd level optimization measure.*
- v. *Assess the induced voltage excitation signal from the parametric chirp input signal.*
- vi. *Design discrete system model for design points input space description.*
- vii. *Perform amplitude modulation as a 2nd level optimization for the PRBS signal, based on online DoE QBC criterion.*
- viii. *Assess the induced voltage excitation signal from the APBRS input signal.*
- ix. *Perform impedance measurements using both amplitude modulation techniques on a solar module.*

1.5.4 QFT signal injection control for the DC-DC converter to ensure accurate real-time measurements of a solar module under variable-point conditions.

These are the detailed objectives to design for and optimise each tier of the control system.

- i. *Characterize the dynamic nonlinear system from the solar module to the converter and load through state space averaging.*
- ii. *Perform small signal analysis to obtain an LTI model for the power optimizer.*
- iii. *Establish control specifications for the power regulation and signal injection control system.*
- iv. *Design lead compensator for the inner current loop and assess its bandwidth performance.*

- v. *Design incremental conductance MPPT algorithm to achieve fast MPP convergence. This algorithm is designed within MATLAB's graphical code interphase.*
- vi. *Quantitatively analyze the PV's output power and battery load profile to identify and characterize the system's expected variations in DC-DC converter.*
- vii. *Quantitatively analyze the PV's bandwidth and impedance parameters to determine the system's expected variations of the ac components necessary for robust signal injection.*
- viii. *Synthesize QFT law for signal injection.*
- ix. *Design power transfer and BIS robust control systems for BIS PV impedance extraction system.*

1.5.5 Feasibility assessment into the use of the SMCs inductor current ripple as an excitation signal.

- i. *Assess the boundary of the converter's switching speed and the impact of varying the switching speed on power regulation.*
- ii. *Formulate design considerations for the inductor current ripple peak-peak magnitude.*
- iii. *Assess design considerations for the inductor current ripple excitation frequency and its correlation with the switching speed required for power regulation.*
- iv. *Analyze the optimal excitation frequency points for the module's impedance parameters.*
- v. *Perform impedance measurements on the solar module and compare the results with an FRA and conventional EIS.*

1.5.6 Scope and limitations of the study

- i. *Only module-level impedance extraction will be performed to demonstrate the ability of the technique to characterize a solar module's AC impedance and hence its operational condition if integrated into a power optimizer – which a module-level device.*
- ii. *Because of the availability of modules, only full cell mono-PERC and polycrystalline solar modules will be studied.*
- iii. *Solar module conditions of assessment will be limited to:*
 - *Power curtailment/operational point variation*
 - *Irradiance variation ($100\text{W}/\text{m}^2 - 800\text{W}/\text{m}^2$)*
 - *Temperature variation ($25^\circ\text{C} - 45^\circ\text{C}$)*
 - *Partial shading*
 - *Hot spots*
 - *Microcracks*

- iv. *IEC-61853-1 standard will be used to select 23 sets of irradiance and temperature points at which to test the operational conditions of the module, as well as the faults diagnostics.*
- v. *Low irradiance conditions were chosen over dark conditions because a DC-DC power converter is used; that is, it requires a certain level of voltage and current to operate. Under dark conditions, the solar module would not produce any power thus, resulting in an inoperable converter. This is because to perform impedance measurements on the solar module, it needs to produce its own power to the converter.*
- vi. *Because of the costs available for the research, microcracks were performed on a flexible polycrystalline solar module; this was to ensure that the primary module used for the experiments was healthy.*
- vii. *System tests will be conducted in a laboratory using solar simulating light bulbs. Due to the limitations in cost, the study will emulate an **80Wp** small scale embedded generation (SSEG) system.*
- viii. *The frequency response analyser has a maximum voltage of **12V** and maximum current of **20A**; this thus, limited the size of the modules to a **12V** module. This does not impact the results because module level impedance measurements are required. This system thus, limited to module level impedance measurements, as well as residential solar systems with storage.*
- ix. *Because AC impedance is being detected, the technique can be scalable to series or parallel connected solar modules. The diodes of the modules are accounted for in the AC impedance measured, while the DC cables only contribute only DC impedance that does not affect the system measurements.*

1.6 Research Outputs/Publications

- i. *L. Shelembe and P. Barendse, "An Adaptive Amplitude-Modulated Pseudo-Random Binary Sequence Excitation for Converter-Based Impedance Spectroscopy Characterization of Photovoltaic Modules," in IEEE Transactions on Industry Applications, vol. 59, no. 2, pp. 2007-2018, March-April 2023, doi: 10.1109/TIA.2022.3218752.*
- ii. *L. Shelembe and P. Barendse, "Inductor current ripple excitation design for rapid converter-based electrical impedance spectroscopy on a monocrystalline solar module," 2022 IEEE Energy Conversion Congress and Exposition (ECCE), Detroit, MI, USA, 2022, pp. 1-8, doi: 10.1109/ECCE50734.2022.9947520.*
- iii. *L. Shelembe and P. Barendse, "Parametric excitation for rapid converter-based PV module broadband impedance estimation," 2022 IEEE Energy Conversion Congress and Exposition (ECCE), Detroit, MI, USA, 2022, pp. 1-7, doi: 10.1109/ECCE50734.2022.9947407.*
- iv. *L. Shelembe and P. Barendse, "An amplitude-modulated pseudo-random binary sequence approach to broadband impedance spectroscopy for photovoltaic module system identification," 2021 IEEE Energy Conversion Congress and Exposition (ECCE), 2021, pp. 443-450, doi: 10.1109/ECCE47101.2021.9595969.*
- v. *L. Shelembe and P. Barendse, "A quantitative feedback theory approach to converter-based broadband impedance spectroscopy for online condition monitoring of photovoltaic modules," 2021 IEEE Energy Conversion Congress and Exposition (ECCE), 2021, pp. 451-458, doi: 10.1109/ECCE47101.2021.9595386.*

- vi. *L. Shelembe and P. Barendse, "Online condition monitoring of Photovoltaic (PV) cells by implementing electrical impedance spectroscopy using a switch-mode DC-DC converter," 2020 IEEE Energy Conversion Congress and Exposition (ECCE), 2020, pp. 2136-2141, doi: 10.1109/ECCE44975.2020.9235596.*

1.7 Plan of development

The organization of this study is detailed in this section. Chapters 2 through 5 focus on the underlying reviews, analyses, design, and experimental considerations required for the implementation of impedance spectroscopy through a switch-mode converter. From the considerations, Chapters 6 through to 9 are applied chapters that demonstrate the major contributions of this study; this is achieved through advanced techniques which tackle several challenging aspects identified during problem formulation and the experimental results, which have not yet been assessed in research. Because of the large volume of literature reviewed, each chapter will include its own set of referenced work. Each chapter's objective and development are summarized as follows.

1.7.1 Chapter 2

The chapter's main objectives are the review and justification into the use of a silicon wafer-based monocrystalline solar module as the device under test (DUT), as well as the review and justification into LIS as the technique of choice for condition monitoring technique for the solar module.

1.7.2 Chapter 3

As the impedance extraction technique of choice, this chapter's main objectives are to conduct a detailed review of linear impedance spectroscopy and justify the use of switch-mode converters as the excitation sources of choice and broadband signals as excitation signals of choice. The justification into the use of a switch-mode converter as the excitation source was undertaken through a comparative review of the current impedance spectroscopy excitation sources, as well as a review into the current state of the art in SMC architecture.

1.7.3 Chapter 4

The chapter's main objective is the justification and subsequent analysis of a boost switch-mode converter's capability in performing linear impedance spectroscopy simultaneously with load power regulation i.e., online linear impedance spectroscopy. The justification for the choice of switch-mode converter is in the form of a comparative review between common switch-mode architectures, in terms of their input and output current continuity and hence, LTI characteristics. The system overview is presented, preliminary design for the control system and DC-DC SMC are presented. Simulation results were presented to assess the converter's power regulation and signal injection capabilities, as well as a comparative assessment of the excitation signals' excitation characteristics.

1.7.4 Chapter 5

Based on the preliminary design and simulation results in Chapter 4, this chapter's objective is to present and analyze the practical considerations necessary to prepare the final system design for implementing online linear impedance spectroscopy on a solar module through an SMC.

1.7.5 Chapter 6

The objective of the chapter is to demonstrate the effectiveness of variable-point control in the induction of a voltage excitation signal through a nonlinear switch-mode converter; this demonstration will be achieved through the parametric uncertainty analysis for both power regulation and signal injection using quantitative feedback theory excitation.

1.7.6 Chapter 7

The objective of the study will be to demonstrate analog excitation signal amplitude modulation on a chirp parametric signal – to be demonstrated through frequency domain convolution; this will be undertaken to mitigate frequency induced nonlinearities, at low and high frequencies, resulting from the converter's dynamic response.

1.7.7 Chapter 8

The objective of the chapter will be the demonstration of digital excitation signal amplitude modulation on a pseudorandom binary sequence. The parametric function design points were synthesized using an adaptive model-based online DoE. The input space for signal amplitude optimization was defined based on the non-linear relationship between the converter APRBS modulating duty cycle and APRBS voltage excitation signal. The APRBS will produce faster impedance estimation results, but with better frequency-induced system non-linearity mitigation and ensure accurate impedance estimation over a wide range of system operating points.

1.7.8 Chapter 9

The objective of the chapter will be the evaluation of a boost converter's inductor current ripple to achieve module triangular excitation; this will be undertaken to reduce the cost associated with processing impedance information on a wide bandwidth solar module.

1.7.9 Chapter 10

Based on the results and analyses, this chapter draws suitable conclusions from the various themes explored in this study; it then makes sensible recommendations for this study and future work.

1.7.10 Appendices

This section presents the appendices which include the laboratory setup and the text-based code used for running the study's algorithms.

1.8 References

- [1] Renewable Energy Policy Network for the 21st Century, p. 31. (2019)
- [2] I. E. Agency, "Snapshot of The Global PV Markets In 2019," International Energy Agency Photovoltaic Power Systems Programme, 2019.
- [3] Federal Energy Regulatory Commission, "Energy infrastructure update for August 2021," August 4, 2021.
- [4] P. Bojek, H. Bahar, "Solar PV", IEA Reports, November 2021
- [5] D. A. Jäger-Waldau, "Importance of operating & maintenance costs (OMCs) relative to overall cost and other cost drivers for PV plants," in Conference Technical Programme Chairman.
- [6] F. O. Hocaoglu et al., "Comparison of experimentally obtained I-V curves of different PV modules," 2018 9th International Renewable Energy Congress (IREC), 2018, pp. 1-4, doi: 10.1109/IREC.2018.8362472.
- [7] H. Denio, "Aerial solar Thermography and condition monitoring of photovoltaic systems," 2012 38th IEEE Photovoltaic Specialists Conference, 2012, pp. 000613-000618, doi: 10.1109/PVSC.2012.6317686.
- [8] T. Fuyuki, T. Tomimoto, A. Tani and Y. Ishikawa, "On-site detection of defective panels in mega solar systems by handy electroluminescence surveillance," 2015 IEEE 42nd Photovoltaic Specialist Conference (PVSC), 2015, pp. 1-3, doi: 10.1109/PVSC.2015.7355838.
- [9] Noboru Katayama, Suguru Osawa, Shunya Matsumoto, Takuma Nakano, Mutsumi Sugiyama, "Degradation and fault diagnosis of photovoltaic cells using impedance spectroscopy", Solar Energy Materials and Solar Cells, Volume 194, 2019, Pages 130-136, ISSN 0927-0248.
- [10] R. A. Guejia-Burbano and G. Petrone, "Partial shading detection on PV panels through on-line Impedance Spectroscopy," *2021 IEEE International Conference on Environment and Electrical Engineering and 2021 IEEE Industrial and Commercial Power Systems Europe (EEEIC / I&CPS Europe)*, 2021, pp. 1-6, doi: 10.1109/EEEIC/ICPSEurope51590.2021.9584621.
- [11] M. I. Oprea et al., "Detection of potential induced degradation in c-Si PV modules using electrical impedance spectroscopy," 2016 IEEE 43rd Photovoltaic Specialists Conference (PVSC), 2016, pp. 1575-1579, doi: 10.1109/PVSC.2016.7749885.
- [12] O. Olayiwola. P. Barendse, "Characterization of photovoltaic cells using broadband impedance spectroscopy," in *Transactions of power electronics*, 2018.
- [13] O. Olayiwola, P. Barendse, "Characterization of Silicon based photovoltaic cells using broadband impedance spectroscopy," in *IEEE Energy Conversion Congress and Exposition (ECCE)*, USA, 2016.

- [15] Koji Takechi and Takeshi Yokoi, "An Impedance Measurement Method of Solar modules by Power Converter using PRBS," in *IEEE Transactions on Industry Applications*, 140(3):140-147 DOI:10.1541/ieejias.140.140
- [16] S. Moore and P. Barendse, "Online condition monitoring of lithium-ion batteries using impedance spectroscopy," in *2017 IEEE Energy Conversion Congress and Exposition (ECCE)*, 2017.
- [17] O. Alao and P. Barendse, "Online Condition Monitoring of Sealed Lead Acid & Lithium Nickel-Cobalt Manganese Oxide Batteries using Broadband Impedance Spectroscopy," in *IEEE Energy Conversion Congress and Exposition (ECCE)*, Portland, 2018.
- [18] O. I. Olayiwola and P. S. Barendse, "Characterization of Silicon-Based Photovoltaic Cells Using Broadband Impedance Spectroscopy," in *IEEE Transactions on Industry Applications*, vol. 54, no. 6, pp. 6309-6319, Nov.-Dec. 2018, doi: 10.1109/TIA.2018.2850025.
- [19] F. A. Aroge and P. S. Barendse, "Time-Frequency Analysis of the Chirp Response for Rapid Electrochemical Impedance Estimation," *2018 IEEE Energy Conversion Congress and Exposition (ECCE)*, Portland, OR, USA, 2018, pp. 2047-2052, doi: 10.1109/ECCE.2018.8558129.
- [20] P. Barendse, O. Faloye, "A Three Level DC-DC Converter for Battery Impedance Spectroscopy," in *Energy conversion conference and exposition*, Baltimore, 2019.
- [21] S. Nejad and D. T. Gladwin, "Online Battery State of Power Prediction Using PRBS and Extended Kalman Filter," *IEEE Transactions on Industrial Electronics*, vol. 67, no. 05, 2020.
- [22] M. T. Talat and A. H. Khawaja, "Li-Ion Battery Parameter Identification Using Pseudo Random Noise," in *2019 15th International Conference on Emerging Technologies (ICET)*, 2019.
- [23] D.-I. S. T. M. R. Jussi Sihvo, "Fast Approach for Battery Impedance Identification Using Pseudo-Random Sequence Signals," *IEEE Transactions on Power Electronics*, vol. 35, no. 03, p. 2548, 2020.
- [24] W. Huang and J. A. Abu Qahouq, "An Online Battery Impedance Measurement Method Using DC-DC Power Converter Control," in *IEEE Transactions on Industrial Electronics*, vol. 61, no. 11, pp. 5987-5995, Nov. 2014, doi: 10.1109/TIE.2014.2311389.
- [25] Z. Xia and J. A. A. Qahouq, "Method for online battery AC impedance spectrum measurement using dc-dc power converter duty-cycle control," *2017 IEEE Applied Power Electronics Conference and Exposition (APEC)*, 2017, pp. 1999-2003, doi: 10.1109/APEC.2017.7930973.
- [26] Y. Cao, H. Yang, T. Gao, S. Shao, and B. Zhang, "Switched-mode Control of Battery Backup Unit in Data Center for Online Impedance Detection," *2020 IEEE Applied Power Electronics Conference and Exposition (APEC)*, 2020, pp. 3332-3336, doi: 10.1109/APEC39645.2020.9124056.
- [27] A. Benshatti, S. M. R. Islam, T. Link, S.-Y. Park, and S. Park, "Design and Control of AC Current Injector for Battery EIS Measurement," in *2020 IEEE Applied Power Electronics Conference and Exposition (APEC)*, 2020.

- [28] H. H. J. W. Jiabin Shen, "Converter-based Electrochemical Impedance Spectroscopy for High-Power Fuel Cell Stacks with Resonant Controllers," *IEEE Transactions on Industrial Electronics*, 2020.
- [29] S. S. Alharbi, S. S. Alharbi, A. M. S. Al-bayati and M. Matin, "A comparative performance evaluation of Si IGBT, SiC JFET, and SiC MOSFET power devices for a non-isolated DC-DC boost converter," 2017 North American Power Symposium (NAPS), Morgantown, WV, USA, 2017, pp. 1-6, doi: 10.1109/NAPS.2017.8107320.
- [30] N. D. Dao and D. -C. Lee, "Passive Soft-Switching Circuit for High Power Density SiC-Based DC-DC Boost Converter," 2020 IEEE Applied Power Electronics Conference and Exposition (APEC), New Orleans, LA, USA, 2020, pp. 2136-2141, doi: 10.1109/APEC39645.2020.9124491.
- [31] S. Tiwari, S. Basu, T. M. Undeland and O. -M. Midtgård, "Efficiency and Conducted EMI Evaluation of a Single-Phase Power Factor Correction Boost Converter Using State-of-the-Art SiC Mosfet and SiC Diode," in *IEEE Transactions on Industry Applications*, vol. 55, no. 6, pp. 7745-7756, Nov.-Dec. 2019, doi: 10.1109/TIA.2019.2919266.

Chapter 2

Photovoltaic module as the device under test

2.1 Introduction

The Sun provides almost all the earth's energy – even fossil fuel energy comes from energy from the sun billions of years ago. The Sun's radiation is made up of a combination of electromagnetic waves having varying wavelengths and energies; radio waves have the lowest energy and highest wavelength, and gamma radiation with the highest energy and smallest wavelength. The Sun's rays contain the full radiation spectrum, with shortwave radiation giving solar cells the capability of producing electricity through the photovoltaic effect; this means that a part of the solar spectrum is useful for generating electricity. Because of diffused light, solar modules have the capacity to function under cloudy days [1] – [3].

The energy supplied by the sun is referred to as the irradiance; it is defined to be the perpendicular flux per unit area of a solar cell's surface and measured in (W/m^2). Outside the earth's atmosphere, this value is approximately, $1367 W/m^2$ [4]. While passing through the earth's atmosphere, the energy and spectral behaviour of the radiation changes; reflections from air particles, clouds, and ground, add diffuse irradiation, which decreases the average global solar irradiance to about $1000 W/m^2$ at sea level on a clear day [5]. The irradiation of the sun can also be affected by the Air Mass (AM), which is defined as the normalized mean length of the solar light's atmospheric trajectory; it describes the attenuation of the irradiance and spectral changes and reduces the power of the light as it gets absorbed by air and dust. The AM outside the atmosphere is (0), at the earth's equator it is (1); and is (1.5) on the earth's surface 48° , with respect to the normal. A PV cell's voltage is significantly dependent on the temperature, as such, solar modules have temperature coefficients, quoted in Kelvin. The coefficients indicate

the rate at which the cell's temperature decreases with increasing temperature. The current and power are both affected by temperature – albeit, not as much as the open circuit voltage. The quoted coefficient for the current is significantly smaller than that of the voltage; reasons for this will be discussed in later sections. One of the major performance characteristics of a solar cell is the fill factor (FF); it is the quotient of the open circuit power and the maximum power for the cell and used as one of the measures of cell efficiency [4] – [6].

Table 2.1 Comparison of different solar cell types

Solar cell type	Efficiency	Advantages	Disadvantages
Monocrystalline	~20%	High efficiency, Optimized for commercial use. High life-time value	Costly Smaller surface area than polycrystalline
Polycrystalline	~15%	Low cost Larger surface area	Sensitive to high temperatures. Lower lifespan Less efficient than mono
Thin Film	~7 – 10%	Low cost Easy to produce. Flexible	Short warranty Short lifespan

Table 2.1 shows that most modern solar cells are made from silicon because of the abundance of the element on the planet. As of 2020, the PV module market was made up of 73% silicon, with the other technologies completing the market [7] – [9]. Crystalline solar cells are shown to be highly efficient at reasonable operational costs. Thin film cells are the cheapest and because of their flexibility, they can be accommodated through more application types compared to the other cells; however, their low efficiency makes them nonideal for high power applications. Thus, silicon-wafer crystalline solar modules were chosen as the modules of choice for characterization in this study because of their wide use in the industry.

2.2 Photovoltaic module electrical characteristics – an overview

This section details the relevant electrical characteristics of silicon-based PV modules. The current and voltage (I/V) curve properties of the module are reviewed in detail, then its power and voltage (P/V) curve is reviewed; lastly, the losses that arise from the module's electrical properties are discussed.

2.2.1 Current and voltage characteristics

The current and voltage characteristics provide a graphical illustration of the relationship between the module's developed potential across its electrodes and the current generated through photovoltaic, under certain conditions of irradiance and temperature. The curve provides crucial information required for the configuration of the module such that it operates at its peak power. A generalized solar I/V curve is illustrated in Figure 2.1.

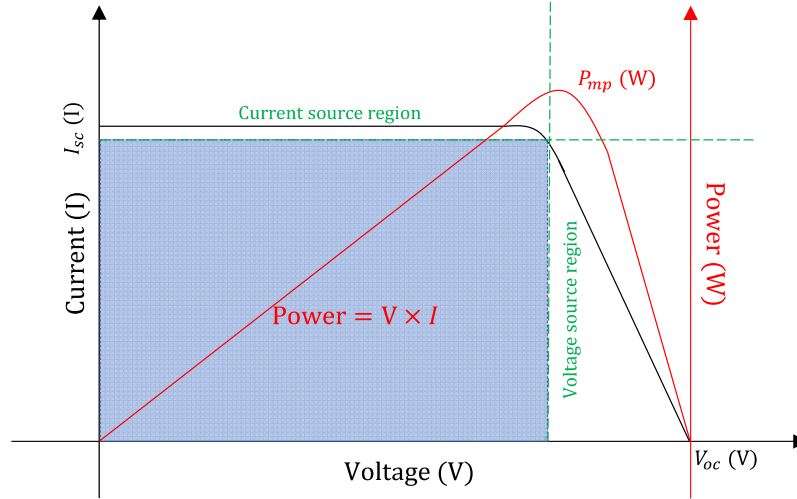


Figure 2.1 Generalised solar module current and voltage characteristics

When the module is short circuited, its photocurrent will be at its maximum – this is the short circuit current, I_{sc} . It is worth noting that this current is higher than the module's current at maximum power, I_{mp} and that the modules voltage will be at its minimum. However, when there is no load connected at the terminals of the module, its potential will be at its maximum i.e., the open circuit voltage, V_{oc} . This voltage is also higher the voltage at maximum power, V_{mp} . The area under the curve of an I/V curve is the power of the power of the module. The point on the curve where area is maximum is called the maximum power point (MPP). Generally, $V_{mp} \cong (80\% - 90\%)V_{oc}$ and $I_{mp} \cong (85\% - 95\%)I_{sc}$. Another important parameter of an I/V curve is the fill factor (FF); it is the ratio of the modules maximum power, $(V_{mp} \times I_{mp})$ and its absolute power, $V_{oc} \times I_{sc}$. The FF is a measure of the quality of the module and is designed to be as close to unity as possible. General values for the FF are between, 0.7 and 0.8 [7] – [11]. The figure illustrates the module's current and voltage source regions. The current source region is defined as such, because in this region, the module provides a steady output current as its potential difference varies. The voltage region is defined as such because the module has a steady potential difference as its output current varies; these regions will be important during the design of impedance measurements excitation signals.

2.2.2 Module internal losses

The chemical properties and physical processes described in the previous sub-sections contribute a significant number of losses within the cell. As of 2022, the maximum efficiency of silicon based solar cells is measured at 25% [12]; this means that most of the sun's energy is not being converted into electrical energy – hence, a closer look at the losses associated with the cells was undertaken.

i. Recombination losses

Generated carriers use emission of energy to separate before they can recombine thus, causing loss of carrier energy and cell performance reduction. There are 4 major contributions to recombination losses that occur at different regions of the cell, the first are surface recombination losses. These losses occur because of the presence of incomplete bonds within silicon atoms; the bonds trap carriers and then causes recombination. The surface recombination velocity is another major contributor to surface recombination losses. This is the velocity at which the generated carriers recombine. It defined as: $S_r = \varepsilon_r v_r N_r$ – where, ε_r is the capture cross section for the holes; v_r is the carrier thermal velocity; and N_r is the number of trapping states; these parameters are generally desired to be low to reduce these losses. The number of trapping states depends on the material – for silicon this value is approximately, $[6 \times 8 \times 10^4]$. Passivation is often used at the surface of the cells to reduce the effect of surface recombination losses [7] – [11].

There are recombination losses that occur in the depletion region – albeit less impactful compared surface recombination losses; this is because of the presence of an electric field in the region. These losses are attributed to the trap assisted recombination or band-band recombination. In the bulk region of the cell, there are significant losses resulting from trap assisted recombination. The present impurities in the cell create an energy state that acts as a trap. Higher purity silicon can drastically reduce the bulk regions recombination losses. The trap assisted bulk recombination can be represented by using the Shockley-Reed Hall effect, U_{shr} .

$$U_{shr} = \frac{pn - n_i^2}{\tau_p(n + n_i) + \tau_n(p + p_i)} \quad (2.1)$$

Where, τ_p and τ_n are the lifetime minority charge carrier in the P and N regions, respectively; n_i and p_i are the equilibrium carrier concentrations. Wafers grown from float-zone have higher minority carrier lifetime, and diffusion constant, D_f – this results in low impurity content in the wafer. The two parameters determine the diffusion length, $L_f = \sqrt{\tau_n D_f}$; this length is defined as the distance the minority carriers need to travel before, they recombine. To ensure low impurity, this length is desired to be as large as possible.

There exist several recombination sites at the contacts of the metal plates. The metal contact junctions are formed both at the back and front of the cell. The back contact junction contributes more to the recombination losses because of its larger contact area. The surface recombination velocity at the back

contacts also increases the losses at these contacts. The light-generated carriers, which are separated by the module's electric field, are collected by the metallic contacts generating the electrical current. If the carriers recombine, there is reduction in the power loss and hence, efficiency [7] – [11].

ii. Series resistance losses

Series resistance losses contribute about 20% of the total losses within a solar cell. This contribution increases significantly when the cell is operated under high intensity. The intensity of the sunlight is measured by the geometrical concentration ratio, γ ; these are proportional to the square of the light intensity of the cells. The losses affect the cell's FF and hence, efficiency. The emitter sheet resistance contributes a large amount of series losses. This resistance is affected by the thickness of the emitter layer; the direction of the current through the emitter; and current collected by the metal fingers. The junction doping concentration is often optimised with a suitable doping concentration to reduce the losses associated with the emitter resistance. Because the junction depth and doping density does not affect the emitter resistance, the depth can be increased to help decrease the sheet resistance [7] – [11].

Much like surface recombination losses, metal semiconductor contact resistance is a major contributor to series resistance losses; the contacts are required for current collection for the cell. There are two types of metal contacts: rectifying and ohmic contacts. Rectifying contacts are rarely used because of their low efficiency; ohmic contacts, however, are used in most modern solar cells because of the better efficiency compared to rectifying contacts – they can either be simple or tunnelling. The nature of the contacts depends on the work function and the electron affinity of the material. Unlike the rectifying contacts, the barrier that forms in the ohmic contacts does affect the electron mobility in any direction. Another way of achieving the contact is by heavily doping the semiconductor – also referred to as tunnelling. This induces surface fields which in turn help reduce the recombination rate at the surface of the metal [7] – [11].

2.3 Photovoltaic module direct current electrical equivalent circuit (DC EEC) models

A photovoltaic cell can be modelled by electrically by studying the internal processes of the cell. Direct current models are large signal models that describe the current-and-voltage curve of PV cells over a range of operational conditions. Because a photovoltaic module is a series combination of photovoltaic cells, a single cell can be studied to analyse module. The DC models are discussed in this section.

2.3.1 Single diode model

The most widely used model for photovoltaic cells is the single diode model as illustrated by Figure 2.2. This model firstly, assumes that the generated current is proportional to the irradiance on the cell and secondly, that the cell is forward bias.

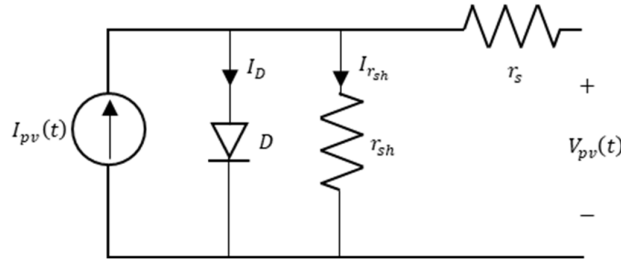


Figure 2.2 Single diode static model for a photovoltaic cell

Where, I_{ph} is the generated photo-current of the cell; I_{shunt} is the current loss due to shunt resistances; I_{diode} is the current loss due to the electron-hole recombination; r_s is the cell's series resistance; and R_{sh} is the shunt resistance of the cell. The parameters define the cell's characteristics for a range of operating regions.

The series resistance is the result of three major causes within the cell: the movement of electrons through the base and emitter of the cell; the resistance that forms between the metal contact and silicon; and lastly, the resistance top and back metal contacts of the cell. The series resistance mainly affects the cell's fill factor, although it can also affect the short circuit current of the cell at high values. The effects of the series resistance at open circuit are null since no current flows through the resistor, but at voltages near open circuit voltage, the resistance can affect the cells operating point. In the presence of the series resistor, the cell's dynamics can be modelled as shown in (2.2). The equation shows that for smaller values of the shunt resistance, the fill factor and hence, cell power is significantly affected. Using Shockley's equation for an ideal diode:

$$I_{pv}(t) = I_{ph}(t) - I_{sat}(t) \left(e^{\frac{V_{pv}(t) + I_{pv}(t)r_s}{nV_T}} - 1 \right) - \left(\frac{V_{pv}(t) + I_{pv}(t)r_{sn}}{r_{shn}} \right) \quad (2.2)$$

Where, where n , is the diode ideality factor (unitless, usually between 1 and 2 for a single junction cell), $I_{sat}(t)$ is the saturation current, and V_T is the thermal voltage, r_{sn} is the normalized series resistance, and r_{shn} is the cell's normalized shunt resistance. By considering effects of the series and shunt resistances, the solar cell can be described fully by equation (2.2) [13][14].

2.3.2 Double diode solar cell model

This model is typically used accurately model polycrystalline cells, which require a model with two diodes, each with different ideal factors; this model is illustrated by Figure 2.3.

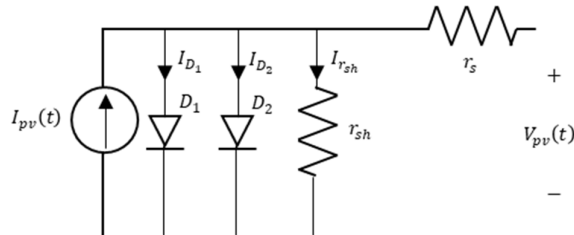


Figure 2.3 Double diode static model for a polycrystalline cell

Diode current, I_{D1} models the diffusion current from the base of the cell to its emitter; while diode current I_{D2} generation-recombination current in the junction space-charge region. However, in wafer-based silicon cells the recombination current is considered negligible apart from low cost mass-produce cells. The model can be mathematically characterized as shown in (2.3).

$$I_{pv}(t) = I_{ph}(t) - I_{ddsat}(t) \left(e^{\frac{V_{pv}(t) + I_{pv}(t)r_s}{V_T}} - 1 \right) - I_{drsatsat}(t) \left(e^{\frac{V_{pv}(t) + I_{pv}(t)r_s}{nV_T}} - 1 \right) - \left(\frac{V_{pv}(t) + I_{pv}(t)r_{sn}}{r_{shn}} \right) \quad (2.3)$$

Where, $I_{ddsat}(t)$ represents the saturation current density resulting from the thermal generation of minority carriers; and $I_{drsatsat}(t)$ represents the saturation current resulting the generation recombination [13].

2.3.3 Reverse bias model

The single diode only considers the cell in its forward bias state, this model looks at the cell's reverse bias characteristics. These reverse bias characteristics are typically neglected since this operation is not the cell's intended design. However, during periods where one cell is partially shaded, it becomes reverse bias and acts a load. Figure 2.4 shows the reverse bias model considering the avalanche breakdown effect.

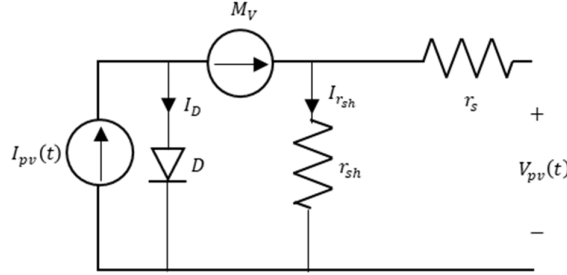


Figure 2.4 Static solar cell reverse bias model considering the avalanche effect.

The figure can be mathematically modelled such that the cell's output current is as represented by (2.4).

$$I_{pv}(t) = \left(1 - \frac{V_{pv}(t)}{V_{br}(t)} \right)^{-m} \left[I_{ph}(t) - I_{sat}(t) \left(e^{\frac{V_{pv}(t) + I_{pv}(t)r_s}{nV_T}} - 1 \right) \right] - \left(\frac{V_{pv}(t) + I_{pv}(t)r_{sn}}{r_{shn}} \right) \quad (2.4)$$

Where, $V_{br}(t)$ is the cell's breakdown voltage in reverse bias mode; and m is the Miller constant. This model can represent the cell's mismatch between a string of cells in a module [13].

2.3.4 Reverse bias model with dynamic effects

While the single diode model and the reverse bias model in the previous section accurately model the cell's characteristics under forward bias and static conditions, they do not consider the cell's dynamic response, nor conditions under which the cell is reverse bias. The dynamic response of cell as illustrated by Figure 2.5 arises when the cell experiences sudden transients.

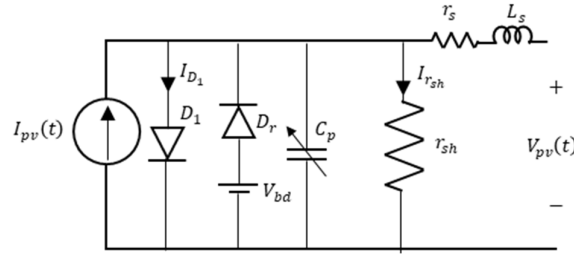


Figure 2.5 PV cell dynamic reverse bias model

Where, C_p is the variable dynamic capacitive impedance in reverse bias; and L_p is the dynamic series inductance. The model above assumes that the irradiance and temperature have no effect on the derived parameters. The saturation current for the forward bias diode is governed by:

$$I_{df}(t) = \frac{I_{sc}(t) + K_i(T - T_n)}{e^{\left(\frac{V_{oc}(t) + K_i(T - T_n)}{nV_T}\right)} - 1} \quad (2.5)$$

Where, K_i is the current temperature coefficient; T_n is the nominal Kelvin temperature. The reverse bias diode current, $I_{dr}(t)$ can be represented as:

$$I_{dr}(t) = I_{sr}(t) e^{\left(\frac{K_r V_{br}(t)}{nV_T}\right)} \left[e^{\left(\frac{K_r V_d(t)}{nV_T}\right)} - 1 \right] \quad (2.6)$$

Where, $V_d(t)$ is the diode's forward voltage drop. The model's dynamic capacitance is made up of the sum of 3 non-linear operational point dependent capacitors: the junction, diffusion, and breakdown capacitances. The junction capacitor, C_j results from the charge stored in the depletion region of the P - N junction. The diffusion capacitance, C_d results from the charge stored in the neutral region of the junction, outside the depletion layer; this capacitance is largest when the cell is operating at its maximum current – where the junction carries a large current; it is deemed as negligible when the P - N junction is reverse biased, as well as being directly proportional to the current carried by the forward bias diode. Lastly, the breakdown capacitance, C_b results from the tunnelling of electrons and the Shockley-Read-Hall recombination. The model of the dynamic capacitance as a function of the cell's temperature and forward bias voltage can be represented as:

$$C_p(T, V_d(t)) = \frac{C_{jo}}{\sqrt{1 - \frac{V_d(t)}{N_p \beta_o}}} + \frac{\tau I_{df}(T, V_d(t))}{nV_T} + \frac{\tau_{bd} I_{dr}(T, V_d(t))}{nV_T} \quad (2.7)$$

Where, C_{jo} is the zero-voltage capacitance; β_o is the zero-voltage potential; τ is the cell's lifetime carrier; and τ_{bd} is the cell's lifetime carrier under breakdown conditions [14].

2.4 Photovoltaic module alternating current electrical equivalent circuit (AC EEC) models.

From the previous section, DC models only characterize series and shunt resistances, thus, having reviewed these models and their shortcomings, AC-EEC models were reviewed. AC equivalent circuits use small AC excitation signal to model electrochemical devices. The small signals ensure that the characterization monitors a single operating point for better characterization. The major difference between the two types of models is that AC-EECs provide the dynamic modelling (while also providing the static characterization), which is missing in DC-EECs. Modelling the dynamics of the cell is vital to better understand firstly, the dynamic behaviour of the cell's junction, secondly; the mobility of cells charge carriers and lastly, the cell's response when connected to non-linear switch mode-converters. The other major advantage of AC-EECs is that they are easier to model compared DC-EECs.

2.4.1 First order Randel's circuit model

A Randle's circuit is an electrical circuit that uses resistive, capacitive, and inductive passive elements to represent an electrochemical system's internal behaviour [15]. The first order iteration of the model, as illustrated by, Figure 2.6 consists of an active series electrolyte resistance and parallel combinations of the double-layer diffusion capacitive and ohmic impedances.

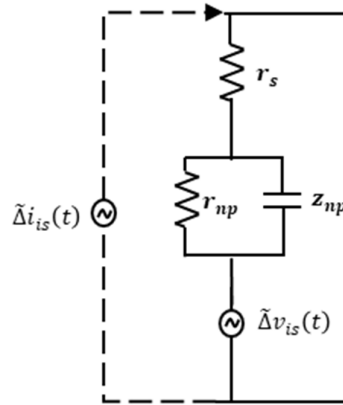


Figure 2.6 First order Randel's AC-EEC model for a photovoltaic cell

From the figure, r_s represents the same series resistance as modelled in the DC-EEC models; the capacitor, c_{np} represents the module's double layer electrolyte-electrode parasitic capacitor, the junction diffusion capacitance; and r_{np} represents the charge transfer resistance and junction diffusion or depletion resistance. The dynamic response can be represented mathematically as illustrated by (2.8):

$$\tilde{\Delta}z_{pv} = r_s + \left[\frac{r_{np}}{(j2\pi f_{is})r_{np}c_{np}} + 1 \right] \quad (2.8)$$

Where, $\tilde{\Delta}z_{pv}$ is the equivalent impedance of the module, and f_{is} is the excitation frequency. The equation for the first order model simpler compared to the single diode model yet, can characterize both static and dynamic behaviour of the cell. The series resistance obtained from an I/V curve results in a single value; however, impedance measurements show that the series resistance is dependent on the module's operating point [15] [16].

2.4.2 Second order Randle's circuit model

While the first order can characterize an electrochemical device, some device dynamics are too fast for the model. Increasing the device excitation frequency, a second order model can be obtained as shown in Figure 2.7.

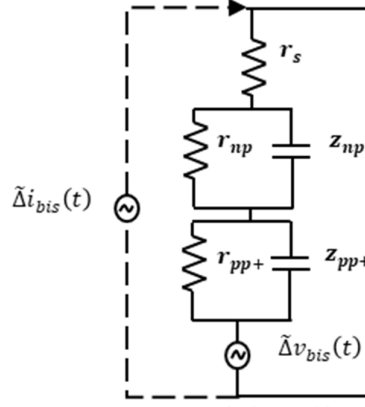


Figure 2.7 Second order Randle's AC-EEC model for a photovoltaic cell

Mathematically, the dynamic response on the figure can be represented as:

$$\tilde{\Delta}z_{pv} = r_s + \left[\frac{r_{np}}{(j2\pi f)r_{np}c_{np} + 1} \right] + \left[\frac{r_{pp+}}{(j2\pi f)r_{pp+}c_{pp+} + 1} \right] \quad (2.9)$$

The second order elements represent the more complex reactions within the cell. From (2.9), r_{pp+} represents the diffusion resistance, while c_{pp+} represents the diffusion capacitance. This model is simplistic but, characterizes the module better than the two-diode DC EEC because it contains dynamic information missing from the two-diode model which only characterizes series and shunt resistance – this provides extra information about the solar PV module. Both c_{np} and c_{pp+} describe the increase in the module's operating voltage resulting from excess minority carriers in the junction. At high operating voltages, the diffusion capacitance increases as a result of high accumulation of charge at the junction. Because capacitance is temperature dependent, the response of the junction capacitance varies with module operating temperature [15] – [17].

2.5 Photovoltaic module faults and degradation review

There are several faults and degradation types that occur on modules for a variety of reasons. As mentioned earlier, silicon wafer-based modules are the most adopted type and as the modules under test in this study, their faults and specific types of degradation will be studied. Only partial shading, micro cracks, and hot spots will be studied, due to the limitations mentioned in the introduction.

2.5.1 Partial shading

Partial shading occurs when one or a group of cells within a module receive less light than the rest of the cells within the module. When this happens, the shaded cells act as loads and excess energy from the fully illuminated cells is dissipated in the shaded cells. They dissipate this energy in the form of heat; this can cause significant power losses for the module. Shading can be caused by numerous factors such as tree shadows, soiling, snow, building structure, etc. On a single cell, the shaded area of the cell is proportional to the amount of power the cell loses. On a module level, the amount of power lost depends on the module configuration and type. The amount of power lost by a module is proportional to the worst shaded cell. If one cell is fully shaded, then the module will produce no power. Manufacturers construct modules such that they can mitigate power losses brought up by partial shading; they do this by employing bypass diodes. The bypass diodes are connected in reverse bias mode between adjacent strings of cells. Generally, each module consists of 3 strings of cells connected in series, with 3 bypass diodes between them. Shading on one or more cells within a string of cells, activates the bypass diodes and the entire string is bypassed. This means that depending on where on the module the shading occurs, the best-case scenario in the event of shading is the loss of one third of the power if one or more cells is partially shaded. With modules assembled with half-cut technology, that loss is only one sixth. Bifacial half-cut modules have even less losses [18] – [20][26].

2.5.2 Micro cracks

Considering that these modules are configured from cells made of silicon – which is a brittle material – they are vulnerable to cracking. These cracks occur on the substrate and are often not detectable by naked eye. Before modules are assembled or installed, these cracks can be caused by numerous factors, including wafer slicing, cell production and the process of stringing and embedding the cell together during module construction. Because some of these cracks are not easily detectable during their early stages, modules are often assembled with cracked cells. During cell production, cells cracks located along the interconnecting ribbon can be caused by the soldering process; there are also cross-cracks caused by needles pressing the wafer; and lastly, cracks appearing at the edge of the cell are caused by heavy objects applying pressure on the cell [22][23].

Packaging and transportation of the modules is another source of cell cracks. It is particularly a significant problem since, in the field, modules aren't subjected to the type of scrutinous testing as they would be during manufacturing. Cell cracks can also be caused by heavy objects landing on the modules

such as tree branches and flying debris brought on by high winds. Cracked cells, typically don't contribute to power losses at their early stages, however, research has shown that this fault manifests in as much as 10% of power losses in after the 6th year of its lifespan [24] – [26].

2.5.3 Hot spots

Burn marks occur when parts of the module experience increased temperatures because of failures in the solar bonding, a break in the connecting ribbon, reverse current flow, or other hot spots. Failures in the solder bonding and ribbon are caused by increased resistance and the associated increase in heat at the joint while large current flow through it. If this continues, the increase in temperature and resistance can cause both the front and back panels of the modules to become discoloured. When a cell becomes reverse bias because of partial shading, cell cracks or defects causing shunt branches, burn marks can appear in severe forms. Burn marks can be associated with large power losses and increased series resistance [18][25][26].

2.6 Photovoltaic module diagnostics tests – a literature review

Having gone through a detailed review of the common faults within crystalline photovoltaic modules, the current state of the art in photovoltaic module diagnostics testing condition monitoring was reviewed.

2.6.1 I/V curve testing

Monitoring the terminal current and voltage characteristics of solar modules is the most common and cost-effective technique employed to detect any anomalies in a residential solar system. With this technique, solar modules are monitored either at a module-level or string-level. The monitoring software available monitors the DC current (and/or voltage) for each solar PV module, and a decrease in the current or voltage on any of the modules is flagged by the system [27]. Technicians are subsequently sent out to manually test each faulty solar module within the strings, using power meters and visual inspection. If an issue detected in one or more modules cannot be rectified, the module(s) is replaced. This type of monitoring is advantageous because it does not require special equipment, it is cheap and scalable to larger systems and because it offers continuous monitoring, in case of a fault, only a small amount of power is lost during the downtime. The main disadvantage with type of diagnostics tool is that it cannot provide fault-specific characterization; it only provides a performance decrease but not what specifically causes the decrease. However, it provides the most comprehensive condition monitoring since each module is individually monitored for any anomalies [28].

In residential systems, devices known as power optimizers have been the latest technology employed for I/V curve characteristics of a solar module. Their main function is to optimize the power output of a solar module, especially in poor conditions, by using DC-DC converters to employ maximum power point tracking (MPPT). Each power optimizer is connected to a single solar PV module, thus offering module-level I/V characteristics – see Figure 2.8. DC string combiner boxes are used to combine the DC cables

from each power optimizer and then connect to a central inverter. This information is processed by a monitoring software that processes all the data received from each power optimizer and thus, offering condition monitoring for each solar module in a string [29].

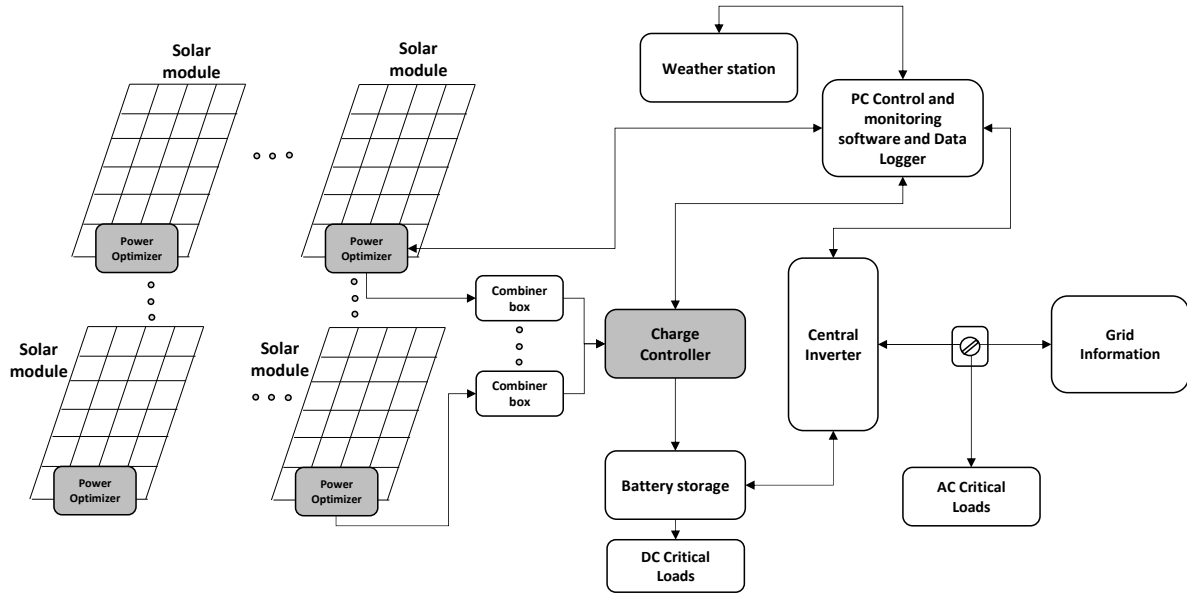


Figure 2.8 Residential solar PV system setup with power optimizers

2.6.2 Radiometric sensors/thermography testing

Infrared thermography is a non-destructive condition monitoring technique that measures the energy radiated from a surface through the analysis of temperature values emitted by the surface. Thermal cameras capture the radiation and convert it into images with distinct colour scales representing the temperature levels on the surface [30]. This type of monitoring technique is more suited to faults that cause a rise in temperature in the solar modules such as, hotspots and cell cracks. There are several variations in implementing thermography that have been adopted. In [31], an analytical based method with artificial neural network for automatic FDD in solar modules was used to implement thermography in solar modules. The main advantage of this type of monitoring is that if implemented using unmanned aerial vehicles it can be used to cover large areas, as well as detect partial shading. The photos taken can also be used to identify larger cracks on the modules. Thus, unlike I/V curves, it provides fault-specific diagnostics but lacks continuous monitoring capabilities.

2.6.3 Electroluminescence testing

This technique is non-destructive, producing image integration measurements as fast as 1s in some applications to detect, mainly, microcracks and other manufacturing defects. In electroluminescence testing, current is injected into a solar module; the radiative recombination of carriers then emits light. In silicon semiconductors most of the recombination occurs because of defects or Auger recombination; band to band recombination which produces radiative emission is low [32].

In [33], a quantitative electroluminescence approach was undertaken for condition monitoring of p-type solar modules influenced by potential induced degradation. To achieve robust results, electroluminescence images of the module were corrected for any camera or perspective related distortion. The relative power loss map was then calculated from the logarithmic ratio of 2 different electroluminescent solar images comparing scenarios either before or after the effects of potential induced degradation. For the quantitative electroluminescence, only average power loss maps were considered, with the highest average at each string evaluated. The results from 2000 solar electroluminescence images and their respective I/V curves showed relative RMS errors between 1% and 3% could be achieved using the technique.

In [34], the performance correlation in field solar modules between the series resistance and electroluminescence imaging was analysed. The 10-year effects of solar bond degradation were analysed through the dark I/V curves were correlated with electroluminescence imaging. The assessment of each cell was carried out by cutting small segments of the frame's back-sheeting and soldering at the cell interconnect. The distribution of the series resistance from the I/V curves showed good correlation with the electroluminescence imaging. In [35], a low-cost electroluminescence imaging approach for automated defect characterization, on solar modules, using modified camera and auxiliary hardware to automate electroluminescence imaging, was undertaken. The camera module was automated using a low-cost Raspberry Pi and a MOSFET. This was shown to achieve characteristics of being portable, affordable, containing high resolution and software extensibility.

2.6.4 Linear Impedance Spectroscopy testing

Linear Impedance Spectroscopy (LIS) is an AC measurement process used to analyse and characterize an electrochemical device's electrical impedance over a range of frequencies. The technique measures either the current or voltage small signal response, from a chosen voltage or current excitation signal. Industry applications and research have shown how LIS – implemented through Frequency Response Analyzers (FRAs) – can be implemented on solar PV modules to measure the impedance – impedance information that can then be used to determine underlying faults such as partial shading, hot spots, cracks, and PID, cell degradation, ageing, mechanical stress – combining a wider range of faults compared to the other techniques discussed [36] – [39]. The results presented in the research showed good impedance estimation, with a significant reduction in cost required for detailed diagnostics, and a reduction in diagnostics acquisition time to minutes, compared to thermography, electroluminescence, or laboratory-based methods – which can take several hours or days in some cases – thus improving the reliability of condition monitoring for these devices [40] – [45].

It is worth noting that, although DC methods can be adopted to characterize electrochemical devices such as batteries, fuel cells and PV modules, AC methods, like LIS, add important information to the EEC diagnostics technique. With AC methods, multiple device parameters can be determined in one measurement; data acquisition time is significantly reduced compared to lab based techniques – which

can take several days – to just minutes using LIS, because the tests are performed at higher frequencies than DC; impedance measurements less than $1\text{m}\Omega$ can be accurately measured; and dynamic device parameters like capacitance and inductance can be measured along with resistive parameters for more detailed diagnostics; this makes frequency based techniques like LIS a superior diagnostics strategy compared to other existing systems, as shown in recent research [46] – [56].

In [38], multi-sine signals were used in the online characterization of silicon-based PV cells. An optimized quasi-logarithmic multi-sine broadband signal was used as the excitation signal of choice. The implementation was achieved through a PEL. The multi-sine signal was optimized in terms of its amplitude to achieve the most optimal crest factor, its frequency spread to optimize the frequency points, and phase to ensure the optimal amplitude profile. The results showed good impedance errors when compared to a commercial frequency response analyser.

In [39], a boost converter was used in the condition monitoring of a solar panel by implementing impedance spectroscopy using a PRBS signal. The PRBS signal was injected through the boost converter's duty ratio while operating in continuous inductor current mode. The PRBS signal was used as a small signal compared to the operating duty ratio signal to ensure linearity within the system. The PRBS signal contains white noise properties and ease of implementation making this the main contribution to the paper. The results were interpreted on a Bode plot assessing normal and partial shading conditions. They showed that the distinguishing factor during partial shading was the increase in the impedance magnitude on a Bode plot.

2.7 Chapter summary

In this chapter, a thorough review of photovoltaic modules was undertaken. The literature showed that poly and monocrystalline silicon photovoltaic modules are the most adopted modules because of their superior efficiency and abundance of silicon, as such, these types of modules were chosen as the devices under test. For electrical diagnostics of a solar module, various electrical equivalent circuit models were reviewed; upon review, AC EECs provided extra information to DC EECs, because they presented dynamic elements that can be used to gather more information. Common faults for crystalline modules were reviewed for diagnostics, these were partial shading, which is common in residential areas, hot spots which are common in systems that experience partial shading, and microcracks which tend to show with age. Lastly, several existing diagnostics techniques for solar PV modules were reviewed. I/V curves are a DC technique that used a decrease in power to detect system faults which is the most widely used and cheapest diagnostics technique, however, it cannot detect specific faults; thermography was the technique of choice in thermal inducing faults such as hot spots but requires special equipment and does not offer continuous monitoring; electroluminescence was good for detecting microcracks and manufacturing defects but also requires special equipment as well modules to be taken offline during the measurements; and LIS is a frequency based technique that uses impedance for diagnostics. The technique can measure a wider variety of faults compared to other reviewed techniques –

faults such as PID, Partial shading, hotspots, cell cracks, degradation, ageing, mechanical stress; LIS does this by characterizing the series resistance as well as capacitances and inductive properties at lower and higher frequencies, thus offering better insight into the properties of the cell. The technique does, however, require linearity in its measurements, by employing small current and voltage signal; this generally requires signal processing units capable of generating such small signals. However, LIS was still a better choice for detailed module-level online diagnostics of solar modules over other techniques, as well as being cheaper, scalable, and non-intrusive thus, the technique of choice and will form the basis for diagnostics of a solar module in this study.

2.8 References

- [1] Kong, J. A. (2008). *Electromagnetic wave theory*. Cambridge, Massachusetts, USA: EMW Publishing
- [2] Shevgaonkar, R. K. (2005). *Electromagnetic Waves*. New Delhi: Tata McGraw-Hill Publishing Company Limited
- [3] Dobbs, R. (1985). *Electromagnetic Waves*. London, England: Routledge and Kegan Paul.
- [4] Myers, D. R. (2017). *Solar Radiation: Practical Modeling for Renewable Energy Applications*. Boca Raton: Taylor and Francis Group.
- [5] Wald, L. (2021). *Fundamentals of Solar Radiation*. London, England: Taylor and Fancis Group.
- [6] Iqbal, M. (2012). *An Introduction to Solar Radiation*. British Columbia, Canada: Academic Press Canada.
- [7] Arnulf Jäger-Waldau, I. f. (2011). *PV Status Report 2011: Research, Solar Cell Production and Market Implementation of Photovoltaics*. Publications Office of the European Union, 2011.
- [8] Cseh, A. (2005). *Market Analysis and Investment Opportunities in Start-up Companies Active in the field of organic photovoltaics*. Norderstedt: Grin Verlag.
- [9] Chowdhury, Md & Rahman, Kazi Sajedur & Chowdhury, Tanjia & Nuthammachot, Narissara & Techato, Kuaanan & Akhtaruzzaman, Md & Tiong, Sieh Kiong & Sopian, Kamaruzzaman & Amin, Nowshad. (2020). An overview of solar photovoltaic panels' end-of-life material recycling. *Energy Strategy Reviews*. 27. 100431. 10.1016/j.esr.2019.100431.
- [10] Achuthan-Bhat. (2006). *Fundamentals Of Semicon Dev*. New Delhi: Tata McGraw-Hill Education.
- [11] Kitai, A. (2011). *Principles of Solar Cells, LEDs, and Diodes: The role of the PN junction*. John Wiley & Sons.
- [12] Nadiya Ivanenko (2022). The 30% efficiency barrier of solar panels has been overcome, 2022.

- [13] F. Jingxun, L. Shaowu, Z. Xianping and L. Yan, "Research on Photovoltaic Cell Models: A Review," 2021 33rd Chinese Control and Decision Conference (CCDC), 2021, pp. 1006-1011, doi: 10.1109/CCDC52312.2021.9602658
- [14] K. A. Kim, C. Xu, L. Jin, and P. T. Krein, "A Dynamic Photovoltaic Model Incorporating Capacitive and Reverse-Bias Characteristics," in IEEE Journal of Photovoltaics, vol. 3, no. 4, pp. 1334-1341, Oct. 2013, doi: 10.1109/JPHOTOV.2013.2276483
- [15] O. I. Olayiwola and P. S. Barendse, "Photovoltaic Cell/Module Equivalent Electric Circuit Modeling Using Impedance Spectroscopy," in IEEE Transactions on Industry Applications, vol. 56, no. 2, pp. 1690-1701, March-April 2020, doi: 10.1109/TIA.2019.2958906
- [16] O. I. Olayiwola and P. S. Barendse, "Characterization of Silicon-Based Photovoltaic Cells Using Broadband Impedance Spectroscopy," in IEEE Transactions on Industry Applications, vol. 54, no. 6, pp. 6309-6319, Nov.-Dec. 2018, doi: 10.1109/TIA.2018.2850025
- [17] O. I. Olayiwola and P. S. Barendse, "Dynamic equivalent circuit modelling of polycrystalline silicon photovoltaic cells," 2017 IEEE Energy Conversion Congress and Exposition (ECCE), 2017, pp. 2310-2317, doi: 10.1109/ECCE.2017.8096449
- [18] A. Houssein, N. Héraud, I. Souleiman and G. Pellet, "Monitoring and fault diagnosis of photovoltaic panels," 2010 IEEE International Energy Conference, 2010, pp. 389-394, doi: 10.1109/ENERGYCON.2010.5771711
- [19] Amardeep Chaudhary, S. G. (2015). Effect of Partial Shading on Characteristics of PV module using Simscape. Int. Journal of Engineering Research and Applications
- [20] J. C. Teo, R. H. (2018, June 17). Impact of Partial Shading on the P-V Characteristics and the Maximum Power of a Photovoltaic String. Energies, pp. 2-22
- [21] Mohamad Hossien Nahidan, M. N. (2021). Power Enhancement under Partial Shading Condition Using a Two-Step Optimal PV Array Reconfiguration. International Journal of Photoenergy
- [22] Ivan Mölken, U. A. (2012). Impact of Micro-Cracks on the Degradation of Solar Cell Performance Based on Two-Diode Model Parameters. Energy Procedia, 167-172
- [23] M. Aßmus, S. S. (2012). Performance Requirements of Crack Detection Systems in Silicon Solar Cell Production. Energy Procedia, 147-153.
- [24] S. Kajari-Schröder, I. M. (2012). Criticality of Cracks in PV Modules. Energy Procedia, 658-663.
- [25] Jaeun Kim, M. R.-C. (2021). A Review of the Degradation of Photovoltaic Modules for Life Expectancy. Energies.
- [26] Marc Köntges, G. O.-A. (2017). Assessment of Photovoltaic Module Failures in the Field. International energy agency photovoltaic power systems programme

- [27] L. Min, S. Huicong, M. Gaofeng and Z. Yu, "Method for measuring I-V characteristics of solar cells," 2019 14th IEEE International Conference on Electronic Measurement & Instruments (ICEMI), Changsha, China, 2019, pp. 966-972, doi: 10.1109/ICEMI46757.2019.9101678.
- [28] E. Barrigón, P. Espinet-González, Y. Contreras, L. Barrutia, I. Rey-Stolle and C. Algora, "On the use of I-V curves as a diagnosis tool for proper external quantum efficiency measurements of multijunction solar cells," 2014 IEEE 40th Photovoltaic Specialist Conference (PVSC), Denver, CO, USA, 2014, pp. 3453-3456, doi: 10.1109/PVSC.2014.6925676.
- [29] A. Elmelegi, M. Aly, E. M. Ahmed and M. M. Alhaider, "An Efficient Low-Cost Distributed MPPT Method for Energy Harvesting in Grid-Tied Three-Phase PV Power Optimizers," 2019 21st International Middle East Power Systems Conference (MEPCON), Cairo, Egypt, 2019, pp. 1042-1047, doi: 10.1109/MEPCON47431.2019.9008027.
- [30] Huerta Herraiz A, Pliego Marugán A, García Márquez FP. Photovoltaic plant condition monitoring using thermal images analysis by convolutional neural network-based structure. *Renew Energy*. 2020; 153:334-348.
- [31] Pierdicca R, Malinverni E, Piccinini F, Paolanti M, Felicetti A, Zingaretti P. Deep convolutional neural network for automatic detection of damaged photovoltaic cells. *Int Arch Photogramm Remote Sens Spat Inf Sci - ISPRS Arch*. 201842; XLII-2:893-900
- [32] K. G. Bedrich et al., "Quantitative Electroluminescence Imaging Analysis for Performance Estimation of PID-Influenced PV Modules," in *IEEE Journal of Photovoltaics*, vol. 8, no. 5, pp. 1281-1288, Sept. 2018, doi: 10.1109/JPHOTOV.2018.2846665.
- [33] K. G. Bedrich et al., "Quantitative Electroluminescence Imaging Analysis for Performance Estimation of PID-Influenced PV Modules," in *IEEE Journal of Photovoltaics*, vol. 8, no. 5, pp. 1281-1288, Sept. 2018, doi: 10.1109/JPHOTOV.2018.2846665.
- [34] A. Sinha, V. S. Pavan Buddha, E. J. Schneller, D. J. Colvin, K. O. Davis, and G. TamizhMani, "Solder Bond Degradation of Fielded PV Modules: Correlation between Performance, Series Resistance and Electroluminescence Imaging," 2019 IEEE 46th Photovoltaic Specialists Conference (PVSC), 2019, pp. 2566-2570, doi: 10.1109/PVSC40753.2019.8980678.
- [35] M. Bazzoli, T. J. Silverman, and L. L. Goddard, "Low-cost electroluminescence imaging for automated defect characterization in photovoltaic modules," 2017 IEEE Photonics Conference (IPC), 2017, pp. 481-482, doi: 10.1109/IPCon.2017.8116193.
- [36] P. Barendse, C. de Beer, "Fuel Cell Condition Monitoring Using Optimized Broad Impedance Spectroscopy," *IEEE Transactions on Industrial Electronics*, 2015.
- [37] O. Olayiwola, P. Barendse, "Characterization of Silicon based photovoltaic cells using broadband impedance spectroscopy," in *IEEE Energy Conversion Congress and Exposition (ECCE)*, USA, 2016.

- [38] O. I. Olayiwola and P. S. Barendse, "Characterization of Silicon-Based Photovoltaic Cells Using Broadband Impedance Spectroscopy," in *IEEE Transactions on Industry Applications*, vol. 54, no. 6, pp. 6309-6319, Nov.-Dec. 2018, doi: 10.1109/TIA.2018.2850025.
- [39] Koji Takechi and Takeshi Yokoi, " An Impedance Measurement Method of Solar modules by Power Converter using PRBS," in *IEEE Transactions on Industry Applications*, 140(3):140-147 DOI:10.1541/ieejias.140.140
- [40] Noboru Katayama, Suguru Osawa, Shunya Matsumoto, Takuma Nakano, Mutsumi Sugiyama, "Degradation and fault diagnosis of photovoltaic cells using impedance spectroscopy", *Solar Energy Materials and Solar Cells*, Volume 194, 2019, Pages 130-136, ISSN 0927-0248.
- [41] R. A. Guejia-Burbano and G. Petrone, "Partial shading detection on PV panels through on-line Impedance Spectroscopy," *2021 IEEE International Conference on Environment and Electrical Engineering and 2021 IEEE Industrial and Commercial Power Systems Europe (EEEIC / I&CPS Europe)*, 2021, pp. 1-6, doi: 10.1109/EEEIC/ICPSEurope51590.2021.9584621.
- [42] M. I. Oprea et al., "Detection of potential induced degradation in c-Si PV modules using electrical impedance spectroscopy," *2016 IEEE 43rd Photovoltaic Specialists Conference (PVSC)*, 2016, pp. 1575-1579, doi: 10.1109/PVSC.2016.7749885.
- [43] O. Olayiwola. P. Barendse, "Characterization of photovoltaic cells using broadband impedance spectroscopy," in *Transactions of power electronics*, 2018.
- [44] O. Olayiwola, P. Barendse, "Characterization of Silicon based photovoltaic cells using broadband impedance spectroscopy," in *IEEE Energy Conversion Congress and Exposition (ECCE)*, USA, 2016.
- [45] Koji Takechi and Takeshi Yokoi, " An Impedance Measurement Method of Solar modules by Power Converter using PRBS," in *IEEE Transactions on Industry Applications*, 140(3):140-147 DOI:10.1541/ieejias.140.140
- [46] S. Moore and P. Barendse, "Online condition monitoring of lithium-ion batteries using impedance spectroscopy," in *2017 IEEE Energy Conversion Congress and Exposition (ECCE)*, 2017.
- [47] O. Alao and P. Barendse, ""Online Condition Monitoring of Sealed Lead Acid & Lithium Nickel-Cobalt Manganese Oxide Batteries using Broadband Impedance Spectroscopy,"," in *IEEE Energy Conversion Congress and Exposition (ECCE)*, Portland, 2018.
- [48] O. I. Olayiwola and P. S. Barendse, "Characterization of Silicon-Based Photovoltaic Cells Using Broadband Impedance Spectroscopy," in *IEEE Transactions on Industry Applications*, vol. 54, no. 6, pp. 6309-6319, Nov.-Dec. 2018, doi: 10.1109/TIA.2018.2850025.
- [49] F. A. Aroge and P. S. Barendse, "Time-Frequency Analysis of the Chirp Response for Rapid Electrochemical Impedance Estimation," *2018 IEEE Energy Conversion Congress and Exposition (ECCE)*, Portland, OR, USA, 2018, pp. 2047-2052, doi: 10.1109/ECCE.2018.8558129.

- [50] P. Barendse, O. Faloye, "A Three Level DC-DC Converter for Battery Impedance Spectroscopy," in *Energy conversion conference and exposition*, Baltimore, 2019.
- [51] S. Nejad and D. T. Gladwin, "Online Battery State of Power Prediction Using PRBS and Extended Kalman Filter," *IEEE Transactions on Industrial Electronics*, vol. 67, no. 05, 2020.
- [52] M. T. Talat and A. H. Khawaja, "Li-Ion Battery Parameter Identification Using Pseudo Random Noise," in *2019 15th International Conference on Emerging Technologies (ICET)*, 2019.
- [53] D.-I. S. T. M. R. Jussi Sihvo, "Fast Approach for Battery Impedance Identification Using Pseudo-Random Sequence Signals," *IEEE Transactions on Power Electronics*, vol. 35, no. 03, p. 2548, 2020.
- [54] W. Huang and J. A. Abu Qahouq, "An Online Battery Impedance Measurement Method Using DC-DC Power Converter Control," in *IEEE Transactions on Industrial Electronics*, vol. 61, no. 11, pp. 5987-5995, Nov. 2014, doi: 10.1109/TIE.2014.2311389.
- [55] Z. Xia and J. A. A. Qahouq, "Method for online battery AC impedance spectrum measurement using dc-dc power converter duty-cycle control," *2017 IEEE Applied Power Electronics Conference and Exposition (APEC)*, 2017, pp. 1999-2003, doi: 10.1109/APEC.2017.7930973.
- [56] Y. Cao, H. Yang, T. Gao, S. Shao, and B. Zhang, "Switched-mode Control of Battery Backup Unit in Data Center for Online Impedance Detection," *2020 IEEE Applied Power Electronics Conference and Exposition (APEC)*, 2020, pp. 3332-3336, doi: 10.1109/APEC39645.2020.9124056.

Chapter 3

Linear impedance spectroscopy – a literature review

3.1 Introduction

As concluded in the previous section, linear impedance spectroscopy – a small signal, non-intrusive, and electrochemical technique used to assess the properties of various materials – was the most suitable technique for detailed online diagnostics for a solar module, offering a wide range of fault detection diagnostics, lower cost for detailed diagnostics, and pre-emptive online monitoring. The technique has applications in many disciplines including engineering, chemistry, medicine et. Al; it is used to analyse and characterise a particular system's electrical impedance over a range of frequencies [1] – [4]. This chapter thus, reviews the technique as a characterization tool in detail, and reviews the most suitable implementation of the technique in terms of its excitation signals and excitation sources.

3.2 Technical process high-level overview

Impedance spectroscopy measures either current or voltage small signal responses, from a chosen voltage or current excitation signals. Small signals are used to ensure that the system remains in its linear and time-invariant (LTI) state. The impedance of the device is obtained by taking the quotient of the excitation and response signals. The process can be executed in either potentiostatic or galvanostatic

mode. Galvanostatic mode measures the impedance of an electrochemical device with a constant voltage supply i.e., a voltage source device; these are devices such as batteries and fuel cells. A current control excitation signal is either added to or induced at the device's input current; a response voltage signal is then measured. Potentiostatic mode is used to measure the impedance of devices that have the characteristics of current sources, such as solar modules. In this mode, a voltage-controlled excitation signal is injected by addition or induction at the source's input and then a response current signal is measured [5] – [7]. Solar modules, however, can be considered both current and voltage source devices as previously reviewed, but under constant light, these modules can produce a constant current. Thus, they are typically operated in their constant current region, therefore measurements will be taken in potentiostatic mode for the solar modules [8] – [10]. The technique is made up of three main sections, device under test, the excitation signal, and excitation source – Figure 3.1 illustrates this. An excitation waveform is synthesized at the discretion of the design engineer, to characterize the impedance of a device – this will be elaborated on in later sections. An excitation source is then chosen to inject the excitation signal – the parameter details of which are at the designer's discretion. The crucial role played by the excitation sources will be discussed in later sections as well.

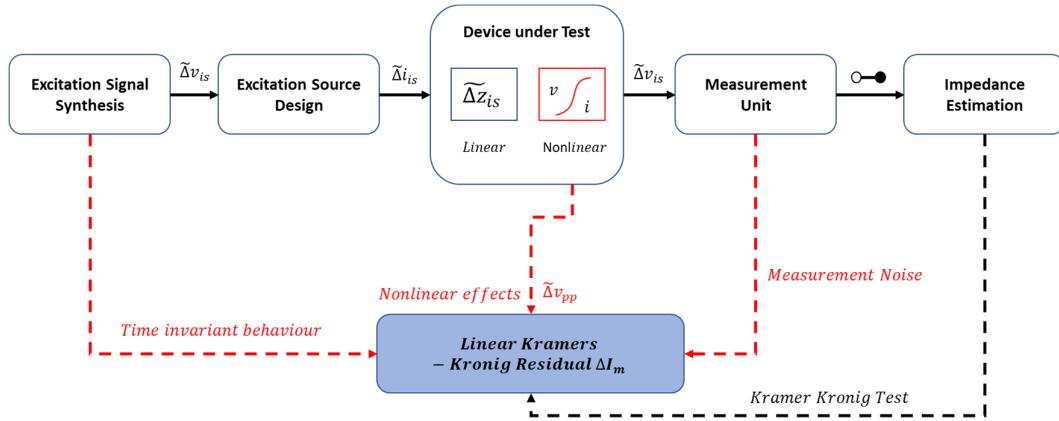


Figure 3.1 Linear impedance spectroscopy high-level process overview

3.3 Impedance representation and curve fitting

The impedance of electrochemical devices is, in almost all cases, complex – meaning it can be decomposed into a complex sum of its real and imaginary components. Equation (3.1) shows the ideal impedance of each of the passive components used in representing an electrochemical device.

$$\left. \begin{aligned} \tilde{Z}_{is,r} &= (r_{is} + 0)\Omega \\ \tilde{Z}_{is,c} &= (0 - j\omega_{is}c_{is})\Omega \\ \tilde{Z}_{is,L} &= (0 + j\omega_{is}L_{is})\Omega \end{aligned} \right\} \quad (3.1)$$

Where, $\tilde{Z}_{is,r}$ is resistive impedance; $\tilde{Z}_{is,c}$ is capacitive impedance; and $\tilde{Z}_{is,L}$ is inductive impedance. It is worth noting that these parameters exist as parasitic passive elements. It is rare that an electrochemical device can be modelled as a purely resistive element. In almost all cases, these devices can be modelled as series and parallel combinations of (3.1). Frequency based responses, such as Bode and Nyquist plots, are used to represent the impedance at various frequencies. Figure 3.2 shows how the impedances are portrayed on a Nyquist plot.

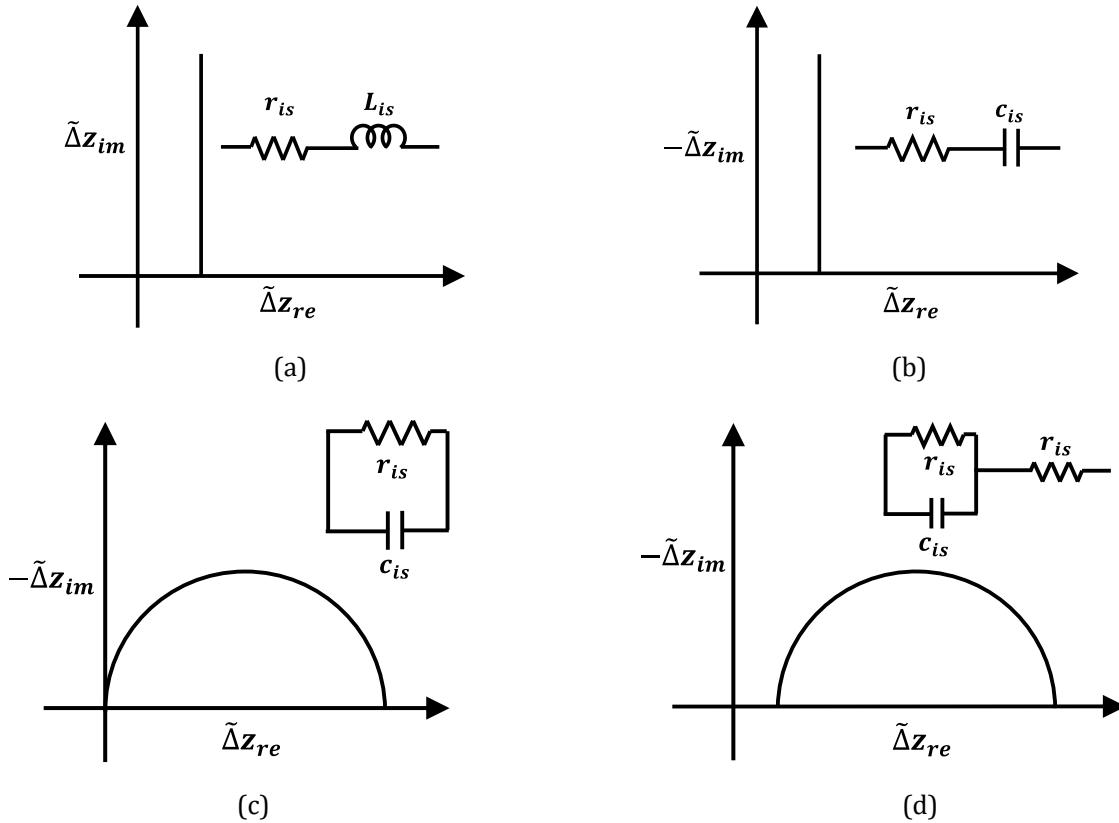


Figure 3.2 Electrical equivalent circuit modelling from impedance spectroscopy data (a) series RL (b) series RC (c) parallel RC (d) series R-RC

Figure 3.2 (a) shows that at low frequencies, the impedance is purely resistive converging to the series resistance value in the RL circuit. As the frequency increases, the impedance diverges towards infinity; this is better explained by (3.1) which show the relationship between frequency and the inductive impedance. Figure 3.2 (b) shows that at low frequencies, the capacitive impedance of the series configuration diverges towards infinity because a capacitor does not conduct DC signals at steady state. At higher frequencies impedance of a capacitor is almost zero and thus, the curve converges to series resistive impedance. Figure 3.2 (c) shows a parallel RC network and at high frequencies the network has very low impedance; this is because, the low impedance of the capacitor at high frequencies short circuits the resistor. At high frequencies, the impedance of the capacitor is almost infinite, as such current would only flow through the resistor; hence the network impedance is purely resistive at low frequencies. Figure 3.2 (d) shows that at high frequencies current would only flow through the series resistor

and the capacitor branch because of the low capacitive impedance at high frequencies. Therefore, the Nyquist plot shows a resistive impedance at high frequencies. At low frequencies the capacitive branch has infinite impedance, thus current would only flow through the resistors, which explains the intersection with the real axis at a value which is the total resistive impedance of the network [11].

3.4 Device excitation signals

This section details the review of impedance excitation signals. There several excitation signals that can be implemented, however, signals that were relevant to the thesis were reviewed.

3.4.1 Introduction

An excitation signal is a waveform synthesized to characterize the behavior of a device through perturbation and then studying, generally, its time or frequency responses. The excitation signal is generally chosen such that measurement noise is minimal while measuring the most accurate response in the shortest possible time. There are 3 main types of noise that are of concern in device excitation, namely, non-coherent, leakage, and nonlinear noise. Non-coherent noise is non-deterministic noise that does not have the same frequency nor a definite phase relationship between 2 waves; nonlinear noise is the result of either transmitter or receiver nonlinearity; and leakage noise is noise that results from choosing a processing unit that does not meet the specifications of either the voltage and current excitations signals chosen. Non-coherent and nonlinear noise are reduced by averaging, while leakage noise is reduced through signal processing and a carefully chosen excitation signal type [12] – [14].

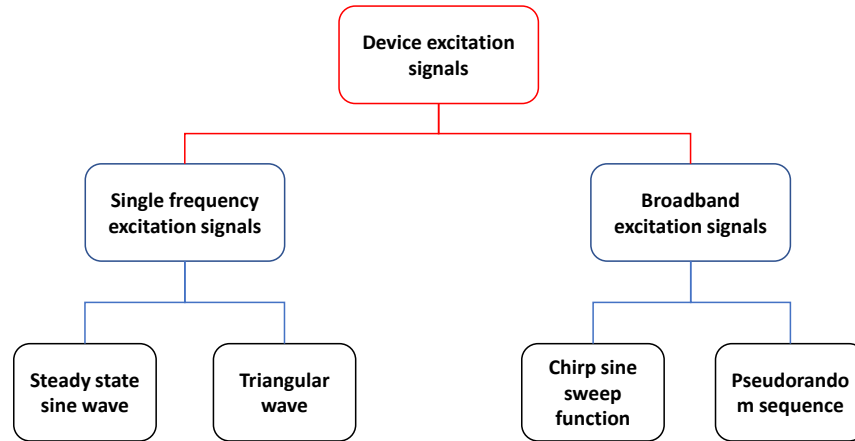


Figure 3.3 Device excitation signal for system identification review

There are several characteristics that govern which excitation signal is chosen for device excitation, these are: RMS-to-peak ratio (crest factor), controlled excitation amplitude, controlled excitation frequency ($-3dB$ cut-off frequency), signal to noise ratio (SNR), test time, leakage minimization and non-linear characteristics. These characteristics are used to benchmark and choose the most optimal excitation signals for a specific DUT excited through a particular type of excitation source. The terminal voltage and current of the solar module can exhibit non-linear internal characteristics. Impedance variation

can be defined for a non-linear system if the requirements of the system's pseudo-linearity are met. Figure 3.3 shows that there are two categories for distinguishing between excitation signals, single-frequency, and broadband frequencies. These are discussed in more detail in the next section.

3.4.2 *Single frequency excitation signals*

These signals are characterized by having a single usable excitation frequency. This means that they do not have a $-3db$ cut-off frequency. They have a varying crest factor depending on the type; they do, however, possess the highest SNR because the signal's power is concentrated at its fundamental frequency. The major drawback of these signals is that they have the longest test time, especially for devices with wider excitation bandwidths. Some common single frequency signals are discussed below.

i. *Single frequency sinusoids*

These signals are most used in impedance spectroscopy because they are simple to implement; implemented practically, they have a high SNR; and they can be implemented on any device and with any excitation source. Because of their simplicity and pure frequency, they provide the most accurate impedance estimation results. Equations (3.2) and (3.3), show the typical voltage excitation and current response signals, respectively with their parameters.

$$\tilde{\Delta}v_{is}(t) = v_m \sin(2\pi f_{is}t) \quad (3.2)$$

Where $\tilde{\Delta}v_{is}$ is the voltage excitation signal; v_m is the controlled excitation amplitude; and f_{is} is the controlled excitation frequency. While the response signal can be described as:

$$\tilde{\Delta}i_{is}(t) = \left(\frac{v_m}{|z_{is}|} \right) \sin(2\pi f_{is}t + \alpha_{is}) \quad (3.3)$$

Where $\tilde{\Delta}i_{is}$ is the device response signal; $|z_{is}|$ is the device's complex impedance magnitude; and α_{is} is the phase difference between the excitation and response signals, resulting from the device's complex impedance. It is important to note that the signal's excitation energy is concentrated at the excitation frequency. All steady state sinusoidal excitation signals have a crest factor of, $\sqrt{2}$ [15]. The results presented in the research showed good impedance estimation, with a significant reduction in cost required for detailed diagnostics, and a reduction in diagnostics acquisition time to minutes, compared to thermography, electroluminescence, or laboratory-based methods – which can take several hours or days in some cases – thus improving the reliability of condition monitoring for these devices [7][15][16].

This is one of the major disadvantages of this because an acquisition time of minutes limits the signal to offline implementation. Figure 3.4 shows that the ideal response of the signal only contains 1 frequency at its excitation fundamental. Although in practice, measurement noise exists within the signal, it falls below the $-3dB$ gain.

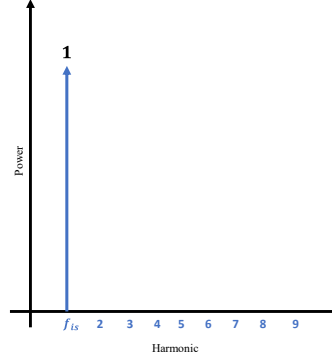


Figure 3.4 Sinusoidal excitation signal ideal frequency response.

In [16], sine wave signals were designed to obtain the frequency response of standard median (SM) and finite impulse response (FIR) median hybrid (FMH) filters. In [5], single sinusoidal signals were used in the system identification of Sealed Lead Acid & Lithium Nickel-Cobalt-Manganese Oxide Batteries through the use of a buck-boost converter. The excitation of these devices showed better performance in terms of accuracy when compared to other excitation signal types. However, the data acquisition time was the longest taking up to 100s, hence, when using sinusoidal signals, accuracy comes at a cost of acquisition time. This acquisition time is not ideal because in 100s there could be a passing cloud or change temperature that compromises the accuracy of the impedance measured. In [7], sinusoidal excitation was used for lithium-ion battery system identification using a bidirectional buck-boost converter. Again, the results showed good accuracy for the sine signals at the cost of longer acquisition times.

ii. ***Triangular wave***

A triangular excitation signal is, in its essence, a definite integral of a square wave excitation signal. As such, its period is half that of either sinusoidal or square wave excitation signal. This can be advantageous because the signal can exist at higher frequencies for similar processing power and sampling capability than both sinusoidal and square wave signals. The crest factor for a symmetrical sine is $\sqrt{3}$, larger than that of sinusoidal waveform. The signal can be described in terms of the Fourier series expansion. For a periodic piecewise continuous function between, $[-T, T]$, its Fourier series expansion is [17]:

$$F_L(t) = x_o + \sum_{i=1}^{\infty} \left[x_i \cos\left(\frac{i\pi t}{T}\right) + y_i \sin\left(\frac{i\pi t}{T}\right) \right] \quad (3.4)$$

where:

$$x_o = \frac{1}{2T} \int_{-T}^T f_L(t) dt \quad (3.5)$$

And for, $i \geq 1$:

$$x_i = \frac{1}{T} \int_{-T}^T f_L(t) dt \cos\left(\frac{i\pi t}{T}\right) \quad \& \quad y_i = \frac{1}{T} \int_{-T}^T f_L(t) dt \sin\left(\frac{i\pi t}{T}\right) \quad (3.6)$$

For a symmetrical voltage excitation signal, the Fourier series expansion becomes:

$$\tilde{\Delta}v_{is}(t) = \left[\frac{1}{2} - \sum_{i=1}^{\infty} \frac{4}{\pi^2 (2\pi i - 1)^2} \cos((2\pi i - 1)\pi t) \right] \forall t \quad (3.7)$$

For a non-symmetrical excitation signal:

$$\tilde{\Delta}v_{is}(t) = \left[\frac{1}{2} - \frac{1}{\pi} \sum_{i=1}^{\infty} \frac{1}{i} \sin\left(\frac{i\pi t}{T}\right) \right] \quad (3.8)$$

The frequency of both spectrums in Figure 3.5 shows a better illustration of the differences in the excitations signal depending on its operating point. When the signal is symmetrical, Figure 3.5 (a) shows that the signal has limited frequency information with odd harmonics, while its power is concentrated at its fundamental frequency. Figure 3.5 (b), however, shows that the non-symmetrical signal is richer in frequency information, and the signal's power is spread out through its response [18].

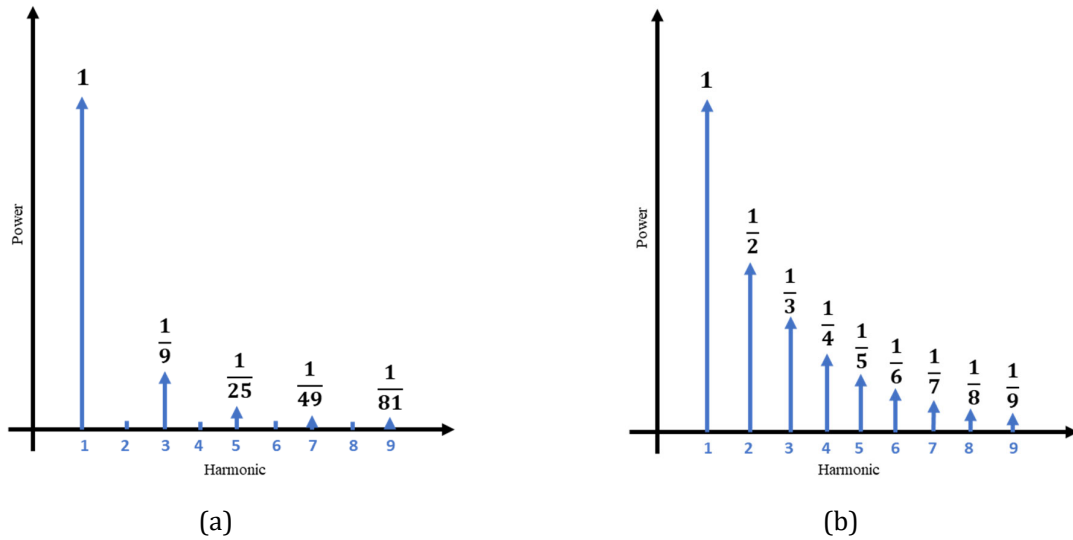


Figure 3.5 Triangular wave excitation signal ideal frequency response (a) 100% symmetry (b) 0% symmetry

In [19], triangular excitation was used as an advanced technique to reduce the data acquisition time in the frequency domain spectroscopy of oil-paper insulation of a transformer, through accelerated dielectric response measurement. Using triangular excitation enabled the measurements to be taken quarter cycles, instead of 2 cycles with the use of square wave excitation. This significantly improved the diagnostics at lower frequencies, which take longer in practice. The technique was verified on 3 different sets of samples and the results showed correlation between the 2-cycle and quarter-cycle measurements, but at a quicker time for the triangular excitation. In [20], triangular excitation was used for

the accelerated sweep frequency response analysis (SFRA) of power transformers. Several triangular waveforms of different symmetry are analysed for the most optimal frequency characteristics. The excitation made use of the fundamental, the second, and third harmonics for characterization. This significantly reduced the data acquisition time for the characterization.

3.4.3 Broadband excitation signals

These signals contain multiple frequencies within the $-3dB$ bandwidth. Their main purpose of their adoption is the reduction of data acquisition time in device system identification. They do this at the cost of accuracy in the measurements compared to single frequency excitation signals. However, this trade-off is often desirable as the margin of error between the single frequency and broadband excitation signals is within acceptable margin of errors. Several common broadband signals were reviewed and discussed below.

i. The chirp signals.

A chirp signal is a sinusoidal frequency sweep analog signal, and as such, the frequency distribution of the signal best illustrates the non-linear effect of the excitation frequency on the induced voltage excitation signal. The standard non-parametric chirp signal can be described mathematically as:

$$\gamma_{chirp}(t) = \cos(2\pi f_{is}(t)t + \omega_o) \quad (3.9)$$

Such that,

$$f_{is}(t) = f_{min} \left(\frac{f_{max}}{f_{min}} \right)^{t-1} \quad (3.10)$$

For the geometric frequency distribution. Where, $f_{is}(t)$ is the excitation frequency of the module; f_{min} is the minimum excitation frequency; and f_{max} is the maximum excitation frequency. It spreads the signal frequencies evenly across a specified bandwidth. This signal only requires frequency distribution optimization because the amplitude is constant for all frequencies. The issue with this signal is that due to the poor signal to noise ratio (SNR) at low frequencies, the low impedance information is scattered. This is solved by implementing either the Short-Time Fourier transform (STFT) or wavelet transforms. The STFT, however, requires optimisation to reduce the signal to noise ratio [21] – [23].

In [24], a comparative study of broadband excitation signals for Li-ion battery impedance spectroscopy, was undertaken. The implementation made use of an STM32 microcontroller as the primary excitation source and signal generator. The results showed that chirp excitation offered excitation energy that was distributed through frequencies of interest. This was, however, only true for a linear chirp

distribution; the discrete Fourier transform (DFT) of the geometric chirp showed an exponentially decreasing amplitude with increasing excitation frequency – this makes the geometric less prone to quantization noise errors at lower frequencies. Therefore, it was shown that chirp excitation is best suited for applications where a dense frequency excitation is required due to its complete harmonic spectrum within its design energy spectrum. In [25], real-time impedance characteristics of an electric energy router were analysed using chirp excitation signals. The excitation signals were injected under closed loop control based on a modulation-demodulation process. An impedance measurement method based on the STFT was proposed. The results showed that with chirp excitation, the impedance of the router could be obtained in real-time with a data acquisition time of 2s. Results also showed with a properly chosen signal injection control with very little dependence on power components, various types of excitation sources could be adopted to implement chirp excitation.

ii. *The Pseudo-random signal*

A PRBS can be used for parameter identification because the frequency spectrum of the signal contains white noise properties. This is a binary periodical signal with an embedded frequency span. Because of the wide impedance bandwidth associated with the PV module, a 12-bit PRBS generator was chosen for better high frequency resolution – as shown in Figure 3.6. Although a higher bit generator could be used for better frequency resolution, this would significantly increase the data acquisition.

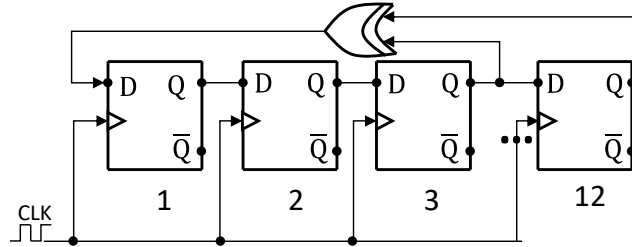


Figure 3.6 Twelve-bit PRBS sequence generator

The excitation signal was optimized in its time response and amplitude. The time domain optimization ensures that the signal has a good usable bandwidth and SNR. The amplitude optimization ensures that the injected duty cycle modulating signal induces a voltage excitation signal that keeps the system within its linear range of operation. The FFT frequency spectrum of an APRBS sequence contains discrete frequencies with a Sinc function envelope as show by Figure 3.7, [26] – [29]. Because of the periodical nature of the signal, the signal power is not concentrated in the impedance frequency band. This is solved by implementing bit-length optimisation. When the signal is optimised, it provides the fastest results because digital signals have faster transients than analog signals.

In [30], PRBS broadband excitation was explored in online impedance measurements of a Li-ion battery. The excitation was used to characterize a Li-ion battery's SoC and SoH. The study presented a systematic approach for designing the PRBS current reference excitation signal for internal measurements of a battery. The results showed that the PBRS excitation signal was able to accurately capture the

internal characteristics of a battery and doing so a fraction of the time it would require analog broad-band techniques. The results did show distortion in the 1Hz components due to the nonlinearity of the battery.

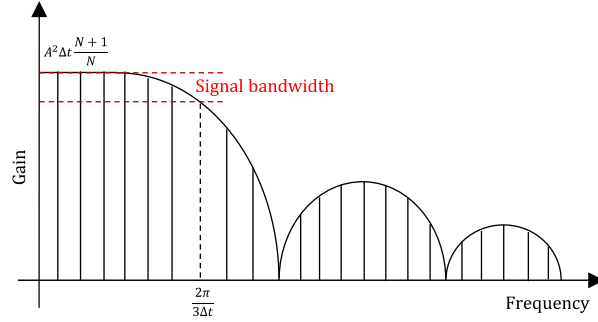


Figure 3.7 Frequency spectrum of PRBS sequence

In [31], converter based PRBS excitation was implemented on solar cells. The impedance measurements of the solar cell were analysed under normal and partially shaded conditions. The impedance measurements of 3 parallel connected solar modules were characterized through the bidirectional boost chopper. The results showed that the impedance measurement value and resonance frequency changes during conditions of partial shading compared to normal conditions, thus showing that the technique and signal can be used to characterize these conditions. The online measurements were compared to the offline measurements and showed accurate correlation, validating both PRBS excitation signal and source for the measurements.

3.5 Excitation sources

There are various excitation sources that can be integrated to implement impedance spectroscopy. A summary of a few is discussed.

3.5.1 AC current generator

For galvanostatic excitation, an AC current generator can be used for device characterization. In galvanostatic excitation, AC current generator is shunted with the DUT, as illustrated by Figure 3.8. The AC current generated is considered the control or excitation signal for the device. Because of the impedance of the device, the AC current generated induces a voltage across the DUT that is considered the response signal. The choice of current generator is dependent upon the application.

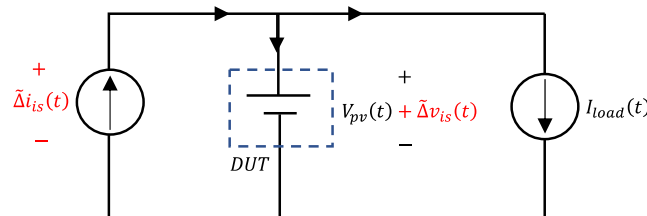


Figure 3.8 AC current generator device excitation

The major drawback to this implementation is that the power required by the current generator to produce the excitation increases, with increasing DUT power; this generally requires additional circuitry for the generator to achieve. The additional circuitry thus increases the system costs and size significantly. The other drawback is that this solution can only be realized through offline implementation [32].

In [33], the design and control of a current injector for battery impedance measurements was undertaken. The AC current generator takes the form of an LC filter-based DC-AC full bridge inverter, while the DUT is a Li-ion battery. The generator measures the impedance of the battery without the need of a battery charger.

3.5.2 Programmable electronics loads

A programmable electronic load (PEL) is an electronic device that can be programmed to emulate various load profiles. In IS, the PEL is used to draw a constant current load with a small AC load component; this small AC load component represents the IS perturbation signal, as illustrated by Figure 3.9.

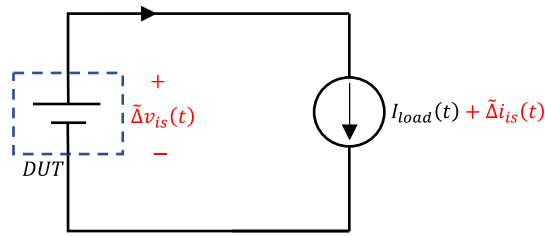


Figure 3.9 Electronic load device excitation

The PEL load draws a programmed load current profile as illustrated by the figure. This forces the DUT to produce a source current proportional to drawn current. This is an example of galvanostatic excitation using the PEL. The advantage of the PEL is that it provides accurate results and can be tailored to almost any desired load profile. The main disadvantage of the PEL is that it only offers offline excitation. Other disadvantages such as, limited dynamic range and options for function generation, high cost, large size, as well as the fact that the DUT needs a power source to work with the PEL, make its adoption difficult in low-cost portable systems [34].

In [35], an electronic load was used in the measurement of a proton-exchange membrane (PEM) fuel cell's internal impedance. The electronic load was designed based on N-MOS technology. The transistor was operated as a variable resistor to support sink the current from the fuel cell. By controlling the gate to source voltage, variable sinking current could be emulated in the transistor. Simulated results on PSPICE from both Bode and Nyquist plots showed that impedance characterization of the fuel cell showed good accuracy for frequencies from 10mHz to 10kHz. In [36], a power electronic load was designed to perform power source characterization. The PEL consisted of 2 stages, an input limitation stage, and an output grid-connection stage. The input limiting stage comprises of a limitation converter designed to assess the power source's performance by limiting both linear and nonlinear loads. The grid

connection stage ensures power balance between the source under test and the grid. The performance of the PEL is analysed based on the Jury test, dynamic factor analysis, and frequency analysis. The results showed that to ensure good load emulation the control scheme and design are crucial in PEL design, and that when designed correctly, converters provide good electronic load emulation.

3.5.3 Switch-mode converters

DC-DC switch-mode converters are electronic devices that are used mainly for DC-DC power regulation. These regulators are essentially nonlinear low pass filters, of at least 2 orders, consisting of at least 2 dynamic storage elements. The major component of the regulator is the power semiconductor switch(s). Pulse width modulating (PWM) techniques are employed in the operation of the switch to ensure good regulation. The modulating signal of the PWM can be designed to achieve a variety of the switching sequences. Depending on the switching sequence chosen, both the input and output can be regulated as desired. Power regulation generally requires a constant modulating signal, provided that there are no system changes. If a small AC modulating signal is superimposed on the modulating signal required for regulation, as shown by Figure 3.10, input current and output voltage signals are induced at the source and load ends, respectively i.e., device excitation signals – this is the basis of performing impedance measurements using an SMC. This, however, is an example of pseudolinear excitation; this is because the modulating must be small enough to ensure that the regulator remains in its linear state [6][7].

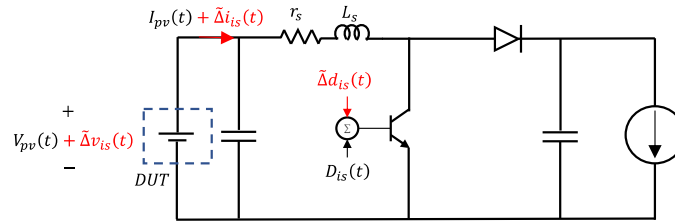


Figure 3.10 Converter device excitation

Because converters employ PWM sequencing, they have the advantage of being feasible for both of-line and online implementation of impedance spectroscopy – depending on the application and need for impedance measurements; this is because the modulating signal responsible for regulating power while the device operates, can be designed to be complex signal, which includes a modulating small AC signal to measure the AC impedance of the device, as discussed earlier. The other major advantages are that regulators are compact and cheaper than commercial PELs and AC generators. However, because the excitation signals are induced through a low pass filtering, pseudolinear device, the accuracy in both the magnitude and phase of the induced excitation signals is compromised. The other disadvantage of employing SMC regulators is that, depending on the limiting inductor size, the input current can have a large high frequency ripple which can push the PV module to a nonlinear state or makes it difficult to measure the response. Filters can be designed to reduce this ripple, but this increases the size and cost

of the system. Selecting a SMC with a small ripple is then desirable. Some regulators also have the disadvantage of a discontinuous input current; this is not desirable because, to perform impedance measurements, the input current required to be continuous to reduce the harmonics at the input such that, the frequencies of interest are easy to isolate, and to observe linearity i.e. only small changes in the operating point during impedance measurements. There have been several advancements in converters – particularly in current ripple reduction – in recent literature, some of the most relevant were reviewed for possible excitation source adoption.

In [37], a boost converter with high gain and zero input current ripple was proposed. The converter features an auxiliary passive clamped network – designed specifically to achieve zero current switching, a soft switching characteristic, and realize continuous zero-ripple current. It also has an asymmetric voltage multiplier cell used to improve the converter’s voltage transfer ratio. Although this seems like an attractive configuration, to achieve the benefits of the converter, as many as 9 storage elements, sensitive to frequency, are required for the converter. Because impedance spectroscopy is a frequency-based technique, the large number of storage elements will make it difficult to predict the effect of the converter’s frequency response on the module’s frequency response. It will also mean that the system is considerable large with significant electromagnetic interference.

In [38], a zero-input current ripple LLC resonant integrated multiphase boost converter was proposed. This proposition was in pursuit of the improvement in power generation efficiency and lifespan in distributed energy devices. The configuration takes the form of a two-stage series connected multiphase boost converter with a full-bridge LLC resonant converter. The converter is able to achieve full soft-switching, that is, zero-voltage and zero-current switching characteristics – this significantly improves the input current ripple during applications of high voltage and low load currents. To better reduce the current ripple, while achieving high gains at duty ratios above 50%, the converter bypasses the use of traditional pulse-width-modulation (PWM) but employs pulse-frequency-modulation (PFM). The drawback to this configuration being the excitation of choice is that it requires 2 power conversion stages through a low frequency transformer. This will have a significant impact on the induced excitation signals, specifically because a transformer makes use of voltage induction as well, hence, the effects of this will be difficult to account for in the impedance measurements.

In [39], a high bandwidth DC-DC switch-mode step down converter with active current ripple injection was proposed. Compared to a conventional buck converter, the proposed converter achieves a high loop bandwidth as well as low input current ripple. The converter is configured with a compensator arranged in a half-bridge with SiC semiconductors, that generates current ripples, for improved efficiency at high power applications. This reduces the size and requirement of passive filters in the converters. The converter can switch between the buck converter’s conventional switches and the compensation network, depending on the required switching requirements of the application. The major issue with this configuration is related to the fact that a buck converter is naturally discontinuous in its input current, meaning that its input current ripple is directly proportional to the module’s output current.

Even though this configuration claims to lower the input current ripple, the discontinuity in the converter still exists, hence, would not be the optimal solution for solar module induced excitation.

In [40], a high-resolution converter for implementing broadband impedance spectroscopy on batteries was proposed. The solution provided considered that when implementing impedance spectroscopy, small current and voltage signals are used, as such, the excitation source used would have to possess the ability to produce both excitation and response signals. A bidirectional buck-boost style converter was used. Impedance spectroscopy was implemented on a lithium-ion battery while the battery charges from an AC generator, such as the grid. This ideal because although the buck converter has discontinuity in its input current, it's output current is continuous. Several resolution improvements were analysed, and the results showed good excitation signal and response signals for the battery. This solution only works for battery applications, however, because these devices can be characterized online through impedance spectroscopy while charging or discharging, whereas solar modules can only be characterized online while modules are producing power. Adopting the solution for solar modules would not work because, although the boost converter has continuous input current, its output current is discontinuous.

3.6 Process LTI conditions

There are several conditions that a design engineer must meet when implementing LIS for device characterization. As mentioned earlier, these conditions are mainly linearity, causality, and time invariance. These concepts are discussed in detail in relation to impedance spectroscopy.

3.6.1 Linearity

Mathematically, linearity is a property of a system which ensures that it follows concepts of addition and homogeneity as shown by Figure 3.11 [41][42]. Electronically, this means that a change voltage or current input leads to a proportional change in its output. This is illustrated by equation (3.11).

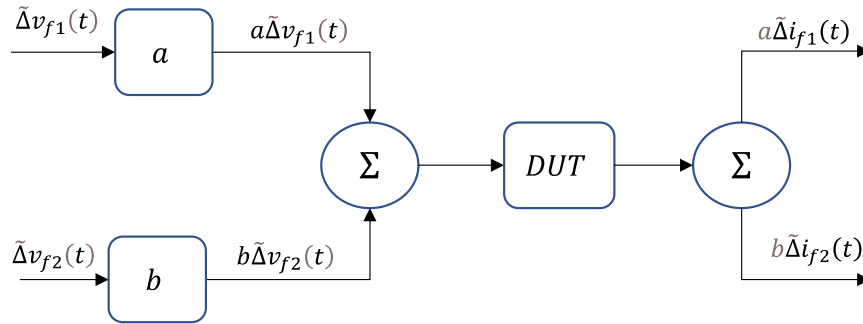


Figure 3.11 Block diagram illustrating linearity of an electronic system

$$a. \tilde{\Delta}v_{f1}(t) + b. \tilde{\Delta}v_{f2}(t)\{DUT\} = a. \tilde{\Delta}i_f\{\tilde{\Delta}v_{f1}(t)\} + b. \tilde{\Delta}i_f\{\tilde{\Delta}v_{f2}(t)\} \quad (3.11)$$

Electrochemical devices operate under specific current and voltage characteristics, as discussed earlier. For a specific current value there will be a specific voltage value – and hence, a specific impedance

value. When either the voltage changes due to load changes or power conditions, the impedance also changes; this means that to accurately characterize the device by studying its impedance, the device's operating point must remain the same in the duration of the process. Because the I/V characteristics of electrochemical devices are not linear within the device's full capacity, impedance spectroscopy is implemented by making use of the concept of pseudolinearity – as illustrated by Figure 3.12 [43]. Pseudolinearity is thus a concept where linearity in a nonlinear device can be assumed by assessing a small section on the device's characteristics.

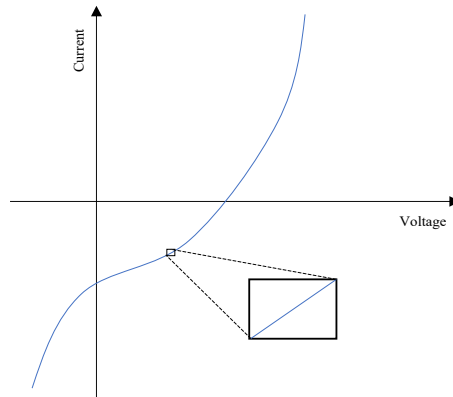


Figure 3.12 Illustration of pseudolinearity from an arbitrary electrochemical device

It is clearly notable that the higher the amplitude, the more likely that the system will operate in a nonlinear region. Figure 3.13 shows that when the system operates in its nonlinear state, the total harmonic distortion (THD) in the response signal is significant. The figure also shows that, although at very small amplitudes the system is linear, the harmonic distortion is significant. This is because at small perturbation amplitudes, the SNR is high. SNR is a measure of how large a signal's power is compared to the noise that the system picks up [44]. With high SNR, the detection of the output response signal is difficult and false harmonics may influence the measured results. The dilemma is thus, the perturbation amplitude must be small enough to keep the system in its linear region, however it must also be large enough to achieve a large SNR for good signal detection. This trade-off is crucial in excitation signal design and its approach is dependent on the type of signal, the linear range of operation of the DUT, and whether the perturbation is direct or induced (more details on this on the next section).

Apart from the perturbation amplitude, the operating point of an electrochemical device plays a vital role in ensuring that measurements are taken in a region that is most linear – this operating point is often referred to as the DC point or DC bias. It is important because these devices are often operated in their saturation region. For example, a battery is discharged at its saturation region such that it can provide a constant potential at its output even with variations in its discharging current. Impedance spectroscopy implemented within the saturation region, however, means that large input perturbation amplitudes are required for relatively small system responses – this then generally leads to nonlinear behaviour observed at the output response signal.

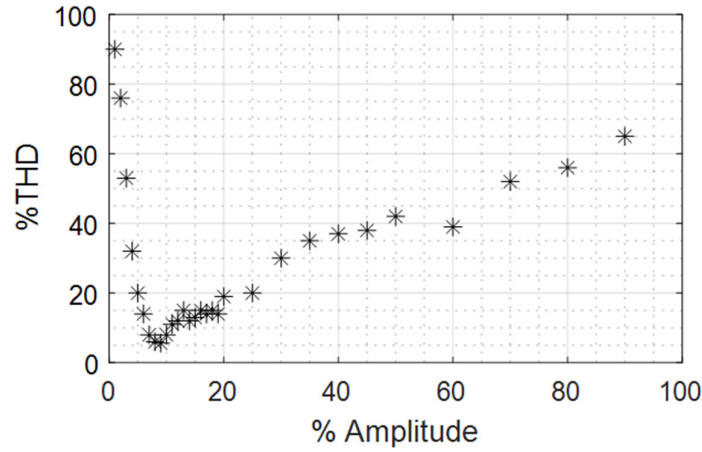


Figure 3.13 Optimal perturbation amplitude for a monocrystalline solar module as a percentage of its DC voltage

In electronic engineering, the linear region of operation in a device refers to region, within a device's characteristics, where an output is directly proportional to changes in its input [44]. Implementing impedance spectroscopy in this region ensures a maximum system response to a perturbation. As discussed in the previous chapter, an AC EEC is approximated using only linear passive components, as opposed to DC EEC approximations, which are based on nonlinear devices such as diodes. This also ensures linearity in the impedance measurements of the device. Ohm's law says that the voltage and current of a conductor are directly proportional, provided the temperature and other physical properties remain the same. Impedance spectroscopy thus characterizes a device by measuring the impedance as its physical properties change – provided that temperature remains the same during the measurements.

3.6.2 Causality

Causality in impedance spectroscopy is a property of a system wherein the output response signal is dependent only on past and present values of the input excitation signal i.e., [42],

$$\tilde{\Delta}v_{f1}(t) + \tilde{\Delta}v_{f1}(t-1) + \dots + \tilde{\Delta}v_{f1}(t-n)\{DUT\} = \tilde{\Delta}i_{f1}(t) + \tilde{\Delta}i_{f1}(t-1) + \dots + \tilde{\Delta}i_{f1}(t-n) \quad (3.12)$$

Where $n \geq 0$. Equation (3.12) means that when the device is perturbed by an excitation signal, the response signal should follow this perturbation. Graphically, the block diagram from Figure 3.14 gives an intuitive representation that simplifies this definition.

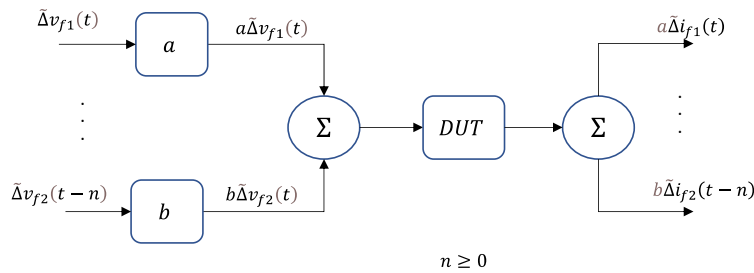


Figure 3.14 Block diagram illustrating causality of an electronic system

3.6.3 Time invariance

Time invariance is the property of a system where its output does not depend on absolute time – as better illustrated by Figure 3.15 [42]. In other words, if a voltage signal, $\tilde{\Delta}v_{is}(t)$ is induced at the input of a device and the system responds through a current signal, $\tilde{\Delta}i_{is}(t)$, then a phase shift, ϕ_{is} in the induced signal will result in a, ϕ_{is} in the response signal, such that:

$$\tilde{\Delta}v_{f1}(t - \phi_{is})\{DUT\} = \tilde{\Delta}i_f(t - \phi_{is}) \quad (3.13)$$

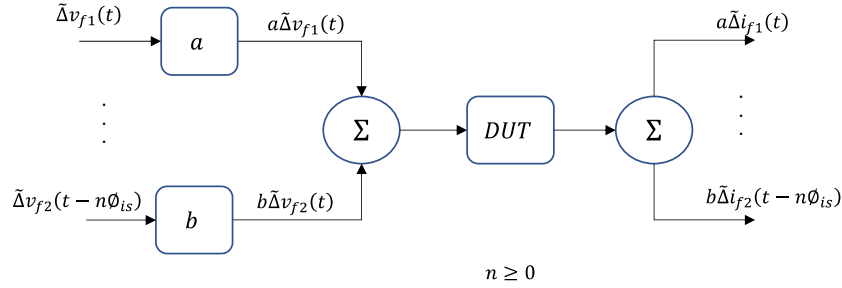


Figure 3.15 Block diagram illustrating time invariance of an electronic system

This property is again, especially crucial in multi-sinusoidal signals. As previously discussed, multiple sinusoidal signals at different phase shifts are used as system inputs; if time invariance holds the response signals at each magnitude will have identical phase shifts to those inputs. It worth noting, however, that there can be phase errors between in the input and output signals because of the presence of phase elements in the whole system; this will be discussed in detail in later sections.

3.7 Impedance spectroscopy – an application review

The current and most relevant literature related to the application of impedance spectroscopy – and its application in relation to solar PV modules and/or cells was reviewed. This was in an effort to understand the current state of the research into implementation of impedance spectroscopy for characterization and diagnostics purposes.

3.7.1 Biomedical applications

In [45], rapid impedance spectroscopy was used for monitoring human tissue impedance, temperature, and treatment outcome, during electroporation-based therapies. These therapies use high voltage pulsed electric fields (PEF), as perturbation signals, to permeabilize tumour tissue, which results in the electrical properties of the tissue, from which, impedance spectroscopy can be used measure these changes. The EIS technique employed in this research aimed to reduce the current standard in data acquisition time of $\sim 10s$, to $\sim 1s$ by making use of inter-burst impedance measurements. The excitation was executed by means of a modified sine wave and an excitation frequency range of between $1.8kHz$

to 4.93MHz. The high frequency impedance measurements enabled the analysis of thermal effects in the measured tissue impedance.

In [46], impedance spectroscopy was implemented during in-vivo measurements of tissue impedance. In-vivo measurements are performed on living tissue – as opposed to ex-vivo measurements. The advantage of implementing impedance spectroscopy on living tissue is that data can be affected by electrode polarization. Single sinusoids were used for excitation at frequencies between 10Hz and 10kHz, while an impedance meter was used as the excitation source of choice. The results showed that it was possible to distinguish between different tissues at high frequencies for modulus and low frequencies for phase. If the detection occurred minutes apart, the results showed that it was possible to compare the impedance of tissues taken either in in or ex-vivo.

3.7.2 Battery applications

In [47], impedance spectroscopy was used in online diagnostics of a lead-acid battery using cross-correlation applied to a boost-type DC-DC power optimizer for batteries. Cross-correlation was implemented as an alternative to competing broadband implementation techniques, but now applied for the purposes of passive network identification. The excitation technique is the same as other converter-based impedance spectroscopy technique; the major change in this implementation is the use of a random integer number white-noise approximation excitation signal – a signal similar to the PRBS with the difference being the optimization parameters and bandwidth.

In [48], instead a PRBS sequence, a ternary sequence was employed in converter-based broadband impedance spectroscopy characterization of a Li-ion battery. The ternary has similar frequency characteristics as a PRBS signals, however, the ternary sequence has twice the power of a PRBS in its frequency spectrum; it also has an intermediary state at “0” giving it 3 states as opposed to 2 states in the PRBS. This gives the signal better properties at dealing with systems that have nonlinearities. The results showed that the ternary sequence performed better at low frequencies where the device experiences difficult transients and nonlinearities and offered the least amount of error when compared to EIS, in the extracted device parameters.

3.7.3 Fuel cell applications

In [49], an open-loop current injection method was used in converter-based EIS characterization for a fuel cell. The proposed system aims to reduce the complexity in the control design of current EIS signal injection techniques by implementing signal injection in open-loop through a multiphase converter with a floating high-voltage end. The amplitude of the injected signals is adjusted adaptively to satisfy the measurement accuracy and linearity condition. The simulation results showed that the induced input current excitation signals have low high-frequency harmonics and within the required linear range.

In [50], the focus was on the online automated analysis of EIS impedance curves for fuel cells using dynamic time warping (DTW). DTW is an algorithm that assesses similarities between two different sequences. The main contribution is the automated analysis of the similarities between impedance curves

for improved diagnostics. The results showed that the method achieved 92% accuracy of correct unsupervised classifications. From the different classes identified, the ohmic resistance is extracted to study the impact of different fuel cell start-up and shut-down protocols.

3.7.4 Solar module applications

In [51], impedance spectroscopy was used in the detection of potential induced degradation of c-Si solar modules under dark conditions. The measurements were performed using an HP 4284A Precision LCR meter at a frequency range of 20Hz to 100kHz. The maximum frequency was chosen based on the fact that above 100kHz, the impedance of the connection circuit becomes dominant. A 2V peak-peak amplitude was used as the perturbation amplitude. The results were compared with electroluminescent and I/V curve results. Impedance spectroscopy method showed good correlation with the other methods, thus confirming the applicability of the technique in assessing PID.

In [52], impedance spectroscopy was implemented on PV modules to characterize partial shading in real-time. The implementation was undertaken using an LTSpice model of the model that included a bypass diode. The research was able to show that through impedance spectroscopy, partial shading on the module could be assessed by only observing the maximum power point. This also indirectly aids MPPT algorithms converge to their global maxima in the presence of mismatches.

In [53], a semiconductor switch approach was undertaken in implementing impedance spectroscopy on a solar module by using a MOSFET as the excitation source. The solar module was used to power the MOSFET through a variable resistive load. The excitation signal is injected by modulating the gate pulses of the switch such that the desired switching sequence is achieved. It should be noted this approach is dynamically similar to that of buck and buck-boost SMC excitation. The results showed good correlation with an FRA when implemented using either single frequency or broadband signals, but best suited for measurements near-open circuit voltage.

3.8 Chapter summary

In this chapter, a thorough analysis of impedance spectroscopy – a frequency-based technique – was undertaken. The high-level process was first presented. The 2 major components of the technique – the excitation signal and source – were then reviewed in detail. The technique's LTI conditions within the context of the DUT of choice were analysed, and lastly, the current state of the art in impedance spectroscopy applications was reviewed and discussed.

The review of device excitation signals showed that single frequency excitation signals offered the best accuracy, ease of implementation and frequency control, whereas the broadband excitation signals offered the quickest data acquisition time and hence, feasibility for online diagnostics and thus the most suitable for the work being proposed.

The review of excitation sources showed that commercial applications make use of FRAs. These devices make use of AC current generators and PELs to sink and source excitation current from a DUT. This

technique offered online offline excitation at a higher accuracy rate, but at higher system size and cost. SMCs were shown to provide both online and offline excitation at lower system costs and lighter in size, however at the cost of reduced accuracy when compared to FRAs. The review into the current research on SMCs showed that there have been several topological advancements that tackle aspects such as, gain, ripple and/or loss reduction. The common theme among these topological changes, however, involves an increase in the number of passive components, such as inductors, capacitors, and semiconductor switches; this is not ideal as increases the overall size and cost of the system, and most importantly, from the perspective of impedance spectroscopy, all the components are frequency responsive. This means that, when attempting to achieve induced excitation of a device through these excitation sources, there will be significant frequency related errors encountered, as well as irregular gain profiles like the peak magnitude response, which could lie within the bandwidth of the module. The phase error will impact the accuracy of the measured impedances, while the irregular gain will impact the linearity of the system, as some frequencies ranges will experience larger or lower gains than other ranges due to the peak response. The stability of a higher order system is more is harder to achieve which might be an issue for applications with varying operating points, as the work presented in this thesis. Hence, the smaller the number of frequency responsive components in the excitation source, the better for the application.

Thus, in conclusion, if the application requires detailed online diagnostics through LIS, broadband signals must be adopted and implemented through an SMC with a minimal number of frequency components, as an excitation source. With this adoption, the impedance of a device under test can be monitored while it executes its conventional application – in the case of solar PV, this would be power regulation. Hence, in applications where a solar PV system is connected to a device that has a DC-DC converter – like power optimizers or charge controllers – impedance spectroscopy can be implemented without the addition of special equipment, interrupting power regulation, or significantly increasing the cost of the system – thus forming the basis for the proposed diagnostics system.

3.9 References

- [1] J. Ross Macdonald, E. B. (2018). Impedance Spectroscopy: Theory, Experiment, and Applications. Florida, US: John Wiley & Sons
- [2] Mark E. Orazem, T. (2017). Electrochemical Impedance Spectroscopy. Florida, US: John Wiley & Sons
- [3] P. Barendse, C. de Beer, "Fuel Cell Condition Monitoring Using Optimized Broad Impedance Spectroscopy," *IEEE Transactions on Industrial Electronics*, 2015.
- [4] O. Olayiwola, P. Barendse, "Characterization of Silicon based photovoltaic cells using broadband impedance spectroscopy," in *IEEE Energy Conversion Congress and Exposition (ECCE)*, USA, 2016.

- [5] O. I. Olayiwola and P. S. Barendse, "Characterization of silicon based photovoltaic cells using broadband impedance spectroscopy," 2016 IEEE Energy Conversion Congress and Exposition (ECCE), 2016, pp. 1-7, doi: 10.1109/ECCE.2016.7854913.
- [6] O. Alao and P. Barendse, "Online Condition Monitoring of Sealed Lead Acid & Lithium Nickel-Cobalt-Manganese Oxide Batteries using Broadband Impedance Spectroscopy," 2018 IEEE Energy Conversion Congress and Exposition (ECCE), 2018, pp. 2026-2032, doi: 10.1109/ECCE.2018.8558352.
- [7] S. Moore and P. Barendse, "Online condition monitoring of lithium-ion batteries using impedance spectroscopy," 2017 IEEE Energy Conversion Congress and Exposition (ECCE), 2017, pp. 5617-5624, doi: 10.1109/ECCE.2017.8096935.
- [8] Dittrich, T. (2018). *Materials Concepts for Solar Cells* (Second Edition).
- [9] Luis Castañer S. (2003). *Modelling Photovoltaic Systems Using PSpice*.
- [10] Rauschenbach, H. S. (2012). *Solar Cell Array Design Handbook: The Principles and Technology of Photovoltaic Energy*. Springer Science & Business Media.
- [11] Engelberg, S. (2005). *A Mathematical Introduction to Control Theory*.
- [12] T. -W. Kang, J. -H. Kim, N. -W. Kang and J. -S. Kang, "A Thermal Noise Measurement System for Noise Temperature Standards in \$W\$ -Band," in *IEEE Transactions on Instrumentation and Measurement*, vol. 64, no. 6, pp. 1741-1747, June 2015, doi: 10.1109/TIM.2015.2398957.
- [13] P. Bareau, A. Abdipour and A. Pacaud, "A new noise measurement method of noise parameters at microwave frequencies," in *IEEE Transactions on Instrumentation and Measurement*, vol. 46, no. 4, pp. 1044-1048, Aug. 1997, doi: 10.1109/19.650824.
- [14] F. Crupi, Gino Giusi, C. Ciofi and C. Pace, "Enhanced sensitivity cross-correlation method for voltage noise measurements," in *IEEE Transactions on Instrumentation and Measurement*, vol. 55, no. 4, pp. 1143-1147, Aug. 2006, doi: 10.1109/TIM.2006.876392.
- [15] N. Chunyan, W. Hao and G. Wenfeng, "Detecting Method of Weak Sine Signal with Initial Phase," 2007 8th International Conference on Electronic Measurement and Instruments, 2007, pp. 3-742-3-745, doi: 10.1109/ICEMI.2007.4351024.
- [16] J. Neejarvi, P. Heinonen and Y. Neuvo, "Sine wave responses of median type filters," 1988., *IEEE International Symposium on Circuits and Systems*, 1988, pp. 1503-1506 vol.2, doi: 10.1109/ISCAS.1988.15215.
- [17] Craig, E. C. (2012). *Electronics via Waveform Analysis*
- [18] Rashid, M. H. (2010). *Power Electronics Handbook: Devices, Circuits and Applications*
- [19] S. Chatterjee, A. K. Pradhan, B. Chatterjee, and S. Chakravorti, "An advanced technique for

- frequency domain spectroscopy of oil-paper insulation at reduced time using triangular excitation," 2015 International Conference on Energy Economics and Environment (ICEEE) 2015, pp. 1-4, doi: 10.1109/Energy Economics. 2015.7235085.
- [20] K. L. I. M. Pramod B. Jayarathna, W. E. P. Sampath Ediriweera, R. Samarasinghe and J. R. Lucas, "Study on Triangular Excitations for Accelerated SFRA of Power Transformers," 2019 9th International Conference on Power and Energy Systems (ICPES), 2019, pp. 1-6, doi: 10.1109/ICPES47639.2019.9105567.
- [21] A. De Angelis, E. Buchicchio, F. Santoni, A. Moschitta and P. Carbone, "Uncertainty Characterization of a Practical System for Broadband Measurement of Battery EIS," in IEEE Transactions on Instrumentation and Measurement, vol. 71, pp. 1-9, 2022, Art no. 1002609, doi: 10.1109/TIM.2022.3156994.
- [22] L. Li, T. Niu, H. Ji, H. Han, and Y. Liu, "Accurate parameter estimation of chirp class signals under low SNR," 2016 IEEE International Conference on Signal and Image Processing (ICSIP), 2016, pp. 412-416, doi: 10.1109/SIPROCESS.2016.7888295.
- [23] J. Speer, K. George, and D. Mutz, "FFT based algorithm to demodulate high frequency chirp signals," 2016 IEEE Aerospace Conference, 2016, pp. 1-6, doi: 10.1109/AERO.2016.7500659.
- [24] A. Fischer, A. Y. Kallel, and O. Kanoun, "Comparative Study of Excitation Signals for Microcontroller-based EIS Measurement on Li-Ion Batteries," 2021 International Workshop on Impedance Spectroscopy (IWIS), 2021, pp. 44-47, doi: 10.1109/IWIS54661021.9711770.
- [25] X. Weng, Z. Zhao, K. Chen, and L. Yuan, "Chirp Signal Injection Method and Real-time Impedance Characteristic Measurement of Electric Energy Router," in IEEE Journal of Emerging and Selected Topics in Power Electronics, doi: 10.1109/JESTPE.2022.3163805.
- [26] H. Lv, J. -C. Fang, J. -X. Xie and P. Qi, "Generating of a nonlinear pseudorandom sequence using linear feedback shift register," 2012 International Conference on ICT Convergence (ICTC), 2012, pp. 432-435, doi: 10.1109/ICTC.2012.6387168.
- [27] Zhigang Cao, Da Fan and Pingyi Fan, "A performance improvement of COFDM system with pseudorandom sequence processing technique," 2005 IEEE International Symposium on Microwave, Antenna, Propagation and EMC Technologies for Wireless Communications, 2005, pp. 1017-1020 Vol. 2, doi: 10.1109/MAPE.2005.1618092.
- [28] Y. Ya-tao, W. Zhi-wei, N. Xin-xin and Y. Yi-xian, "A New Way to Construct Pseudorandom Sequence," 2007 2nd IEEE Conference on Industrial Electronics and Applications, 2007, pp. 2513-2515, doi: 10.1109/ICIEA.2007.4318863.
- [29] D. A. VISAN, I. LITA, M. JURIAN, and M. GHERGHE, "Pseudorandom Sequence Generator for Spread Spectrum Communications," 2018 10th International Conference on Electronics, Computers and Artificial Intelligence (ECAI), 2018, pp. 1-4,

- [30] J. Sihvo, T. Messo, T. Roinila and R. Luhtala, "Online Internal Impedance Measurements of Li-ion Battery Using PRBS Broadband Excitation and Fourier Techniques: Methods and Injection Design," 2018 International Power Electronics Conference (IPEC-Niigata 2018 -ECCE Asia), 2018, pp. 2470-2475, doi: 10.23919/IPEC.2018.8507565.
- [31] T. Yokoi, K. Takechi, and H. Kakigano, "Impedance Measurement Method for Solar Cell Evaluation using a Power Converter," 2019 IEEE Third International Conference on DC Microgrids (ICDCM), 2019, pp. 1-6, doi: 10.1109/ICDCM45535.2019.9232790.
- [32] W. Huang and J. A. Abu Qahouq, "An Online Battery Impedance Measurement Method Using DC-DC Power Converter Control," in IEEE Transactions on Industrial Electronics, vol. 61, no. 11, pp. 5987-5995, Nov. 2014, doi: 10.1109/TIE.2014.2311389.
- [33] A. Benshatti, S. M. R. Islam, T. Link, S. -Y. Park and S. Park, "Design and Control of AC Current Injector for Battery EIS Measurement," 2020 IEEE Applied Power Electronics Conference and Exposition (APEC), 2020, pp. 3452-3455, doi: 10.1109/APEC39645.2020.9124463.
- [34] H. C. Kanchev, "Modeling of Boost Converter-based Electronic Load with Energy Recycling Capability," 2018 IX National Conference with International Participation (ELECTRONICA), 2018, pp. 1-4, doi: 10.1109/ELECTRONICA.2018.8439666.
- [35] A. Bouaicha, H. Allagui, A. Mami and A. Rouane, "Study of an electronic load for measuring the internal impedance of a PEM fuel cell," 10th International Multi-Conferences on Systems, Signals & Devices 2013 (SSD13), 2013, pp. 1-5, doi: 10.1109/SSD.2013.6564146.
- [36] Huang Zhaoxia and Zou Yunping, "Design and analysis on the Power Electronic Load," 2009 IEEE 6th International Power Electronics and Motion Control Conference, 2009, pp. 2649-2653, doi: 10.1109/IPEMC.2009.5157855.
- [37] J. Ai, M. Lin, Y. Kong, and P. Wang, "A Ripple-free Input Current Boost Converter with Asymmetric Voltage Multiplier cell for Distributed PV Generation Systems," 2020 23rd International Conference on Electrical Machines and Systems (ICEMS), 2020, pp. 1321-1325, doi: 10.23919/ICEMS50442.2020.9291106
- [38] Y. Dong, Y. Wang, F. Wang, and S. Wang, "Zero Input Current Ripple LLC Resonant Converter with Integrated Multiphase Boost Converter," 2021 IEEE International Conference on Computer Science, Electronic Information Engineering, and Intelligent Control Technology (CEI), 2021, pp. 602-606, doi: 10.1109/CEI52496.2021.9574556.
- [39] S. Mao, C. Gao, and Y. Lu, "High Bandwidth DC-DC Converter with Active Current Ripples Injection," 2019 22nd International Conference on Electrical Machines and Systems (ICEMS), 2019, pp. 1-5, doi: 10.1109/ICEMS.2019.8921809.
- [40] S. K. Dam and V. John, "High-Resolution Converter for Battery Impedance Spectroscopy," in IEEE Transactions on Industry Applications, vol. 54, no. 2, pp. 1502-1512, March-April 2018,

doi: 10.1109/TIA.2017.2771498.

- [41] O. M. M. Vall, "Mixed sensitivity H_∞ control for LTI systems with varying time delays," in *Journal of Systems Engineering and Electronics*, vol. 29, no. 3, pp. 611-617, June 2018,
- [42] C. D. Pathiranage, K. Watanabe, B. Jayasekara and K. Izumi, "Simultaneous Localization and Mapping: A Pseudolinear Kalman Filter (PLKF) Approach," 2008 4th International Conference on Information and Automation for Sustainability, 2008, pp. 61-66,
- [43] F. Harris and C. Dick, "SNR estimation techniques for low SNR signals," The 15th International Symposium on Wireless Personal Multimedia Communications, 2012, pp. 276-280.
- [44] V. O. Turin, R. S. Shkarlat, G. I. Zebrev, B. Iñiguez and M. S. Shur, "The "Extrinsic" Compact Model of the MOSFET Drain Current Based on a New Interpolation Expression for the Transition Between Linear and Saturation Regimes with a Monotonic Decrease of the Differential Conductance to a Nonzero Value," 2020 4th IEEE Electron Devices Technology & Manufacturing Conference (EDTM), 2020, pp. 1-4, doi: 10.1109/EDTM47692.2020.9117810.
- [45] M. F. Lorenzo, S. P. Bhonsle, C. B. Arena, and R. V. Davalos, "Rapid Impedance Spectroscopy for Monitoring Tissue Impedance, Temperature, and Treatment Outcome During Electroporation-Based Therapies," in *IEEE Transactions on Biomedical Engineering*, vol. 68, no. 5, pp. 1536-1546, May 2021, doi: 10.1109/TBME.2020.3036535.
- [46] D. Meroni, D. Bovio, M. Gualtieri and A. Aliverti, "In-vivo Measurements of Tissue Impedivity by Electrical Impedance Spectroscopy," 2018 40th Annual International Conference of the IEEE Engineering in Medicine and Biology Society (EMBC), 2018, pp. 1-4, doi: 10.1109/EMBC.2018.8513502.
- [47] T. N. Gücin and L. Ovacik, "Online Impedance Measurement of Batteries Using the Cross-Correlation Technique," in *IEEE Transactions on Power Electronics*, vol. 35, no. 4, pp. 4365-4375, April 2020, doi: 10.1109/TPEL.2019.2939269.
- [48] J. Sihvo, D. Stroe, T. Messo and T. Roinila, "Fast Approach for Battery Impedance Identification Using Pseudo-Random Sequence Signals," in *IEEE Transactions on Power Electronics*, vol. 35, no. 3, pp. 2548-2557, March 2020, doi: 10.1109/TPEL.2019.2924286.
- [49] Y. Deng, Z. Yao, G. Luo, M. Liu, and W. Chen, "An Open-loop Current Injection Method Based on Multiphase DC-DC Converter for Fuel Cell Electrochemical Impedance spectroscopy," 2021 IEEE 2nd China International Youth Conference on Electrical Engineering (CIYCEE), 2021, pp. 1-8, doi: 10.1109/CIYCEE53554.2021.9676863.
- [50] M. Durand et al., "Automated Analysis of EIS curves for PEM Fuel Cells using Dynamic Time Warping," 2021 IEEE 13th International Symposium on Diagnostics for Electrical Machines, Power Electronics and Drives (SDEMPED), 2021, pp. 303-309, doi: 10.1109/SDEMPED51010.2021.9605563.

- [51] M. I. Oprea et al., "Detection of potential induced degradation in c-Si PV modules using electrical impedance spectroscopy," 2016 IEEE 43rd Photovoltaic Specialists Conference (PVSC), 2016, pp. 1575-1579, doi: 10.1109/PVSC.2016.7749885.
- [52] R. A. Guejia-Burbano and G. Petrone, "Partial shading detection on PV modules through on-line Impedance Spectroscopy," 2021 IEEE International Conference on Environment and Electrical Engineering and 2021 IEEE Industrial and Commercial Power Systems Europe (EEEIC / I&CPS Europe), 2021, pp. 1-6, doi: 10.1109/EEEIC/ICPSEurope51590.2021.9584621.
- [53] O. I. Olayiwola and P. S. Barendse, "Power Electronic Implementation of Electrochemical Impedance Spectroscopy on Photovoltaic Modules," 2020 IEEE Energy Conversion Congress and Exposition (ECCE), 2020, pp. 3654-3661, doi: 10.1109/ECCE44975.2020.9236404.

Chapter 4

Impedance extraction system design, analysis, and simulation results

4.1 Introduction

As concluded in the previous section, SMCs with the least number of storage elements should provide the most preferred solution to online device characterization through impedance spectroscopy. As such, there are several configurations of such SMCs, each with their advantages and disadvantages for various applications. They can be classified into two or more levels depending on the number of voltage levels that the primary energy storage element of the converter can acquire. The SMCs can also be classified into 1 or more phases depending on the number of parallel branches present at the input of the circuit. Both these two factors affect the input current and output voltage ripples of the converter.

Classical 2-level, single-phase SMCs are the most widely used type because of their simple and compact configurations. There are 3 types of such SMCs, namely, the buck, boost, and buck-boost SMCs; they are configured from a single inductor, output filter capacitor, diode, and semiconductor switch [1]. A buck converter is a step-down SMC that transfers power from a high voltage source to a low voltage load [2]. In the context of BIS applications, it has a series switch at the input, which means that when the switch goes from ON to OFF, both PV module and converter are instantly disconnected i.e., the input current is zero. When the switch goes from OFF to ON, the input current jumps from zero to a high value. This sequence causes the input current waveform to have a large discontinuity, and hence, large current ripple, which is not desirable for LIS, as discussed in the previous section. The other issue is that when this large change in the input current, affects the operating point of the system. As discussed in the

previous chapter, implementation of impedance spectroscopy is achieved through pseudolinearity which relies on small changes in the system, thus the large change in the input current violates the linearity clause. Hence, implementing impedance spectroscopy using the buck converter would require a large input filter to make the input current continuous to alleviate these issues. The buck converter does have continuous load current, thus, making it an ideal excitation source for battery applications – that is, implementing impedance estimation while the battery changes – as opposed to the traditional implementation during its discharging cycle.

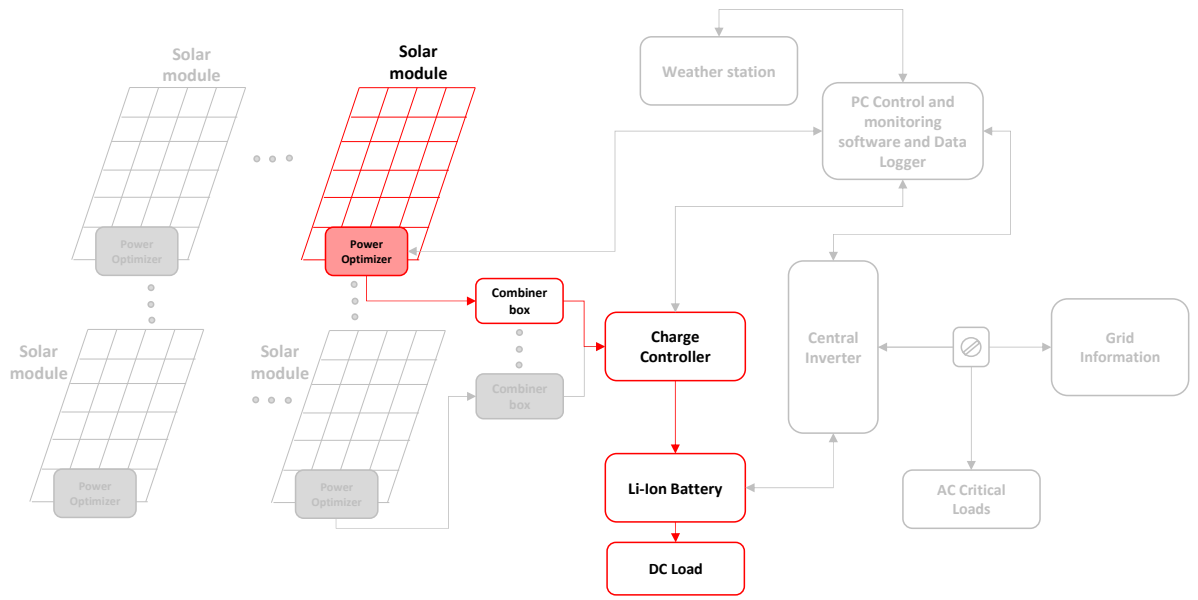
A buck-boost converter is an SMC that transfers power by stepping voltage either up or down, depending on the requirements of the load [3]. It, however, suffers from the same discontinuity as the buck converter due to its switch being in series with the power device – hence, shares the same drawbacks as a buck converter when implementing linear impedance spectroscopy. For a buck-boost converter both its input and load currents are discontinuous, making it a non-ideal excitation source for any applications, unless topological changes are applied to it, or special conditions considered when making use of it.

Lastly, a boost converter is a step-up SMC that transfers power from a low voltage source to a high voltage load. It is configured from a shunt switch and a series inductor with respect to the input source [4]; this means that the input current is continuous, regardless of the state of the switch. Clearly a boost SMC provides the best solution to impedance spectroscopy implementation because it provides continuous input current and hence, the lowest input current ripple. To further reduce the input current ripple of two-level, single-phase boost SMCs, two or more phases, making it a multiphase SMC, can be added in parallel to each configuration [5]. This gives the boost SMC the potential to theoretically reduce the input current ripple to zero, depending on the operating point. The drawbacks of the configuration are the potential mismatch between each of the phases; increased system size and weight; and increased system cost, due to the extra components required. However, the advantages such as, better transient stability, faster transients, better switch utilization, and better efficiency [6], outweigh the drawbacks as they are crucial for impedance spectroscopy. Hence, this configuration will form the basis of the study's excitation source.

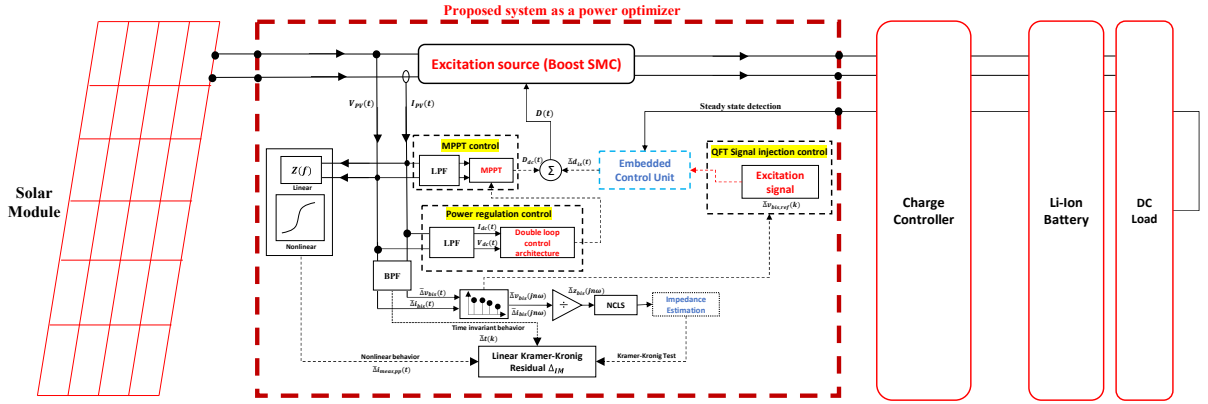
As concluded in the previous section, in applications where a solar PV module is connected to a device with a DC-DC converter – like power optimizers – impedance spectroscopy can be implemented; this is explored in detail.

4.2 Impedance extraction system overview

Figure 4.1 shows the system overview for the proposed impedance extraction system of the solar module; this figure is derived from parts of Figure 2.8 illustrating where a power optimizer fits into the overall system as depicted by Figure 4.1 (a). As illustrated by Figure 4.1 (b), the solar module delivers excess power to the load battery while impedance measurements are performed. The overview of the process is discussed below.



(a)



(b)

Figure 4.1 Proposed system block diagram demonstrated as a power optimizer for online impedance extraction of a solar module (a) Location of the proposed system on a residential PV system (b) Detailed module-level system description.

- During power regulation, the module's voltage and current are measured and a low pass filter (LPF) is used to extract only the DC components, while the AC components are filtered out. The power regulation control algorithm produces voltage setpoint for the MPPT control in order to produce the DC modulating duty cycle, $D_{ac}(t)$ which is responsible for controlling the regulation of DC power transfer from the solar PV module to the load. This is how a conventional power optimizer or charge controller works.
- As an advancement to conventional power optimizers, the figure shows how a single modulating BIS duty cycle, $\Delta d_{is}(t)$ (provided it has a small enough amplitude) added to DC modulating duty cycle, can be used to simultaneously perform impedance measurements and power regulation. During this process, the module's voltage is measured using a band pass filter (BPF) to remove high frequency noise and the inductor current ripple, as well as the DC signals, extracting only the frequencies within the solar module's bandwidth. Within the

power optimizer, an appropriate excitation signal is generated through software, with specifications required to excite the module within its frequency specification. The embedded control unit is then responsible for ensuring that all linearity conditions are met before producing the BIS modulating duty cycle. This duty cycle is then again superimposed (added) on the DC modulating duty cycle to produce the boost SMC's modulating duty cycle, $D(t)$. PWM is then used to produce the pulses necessary for the switch of the SMC.

- At steady state, the BIS modulating duty cycle induces a small AC voltage excitation signal, $\Delta v_{is}(t)$ at the input of the solar module. Signal injection control is responsible for ensuring that the BIS induced voltage signal has an amplitude profile that keeps the solar module operating in its linear range of operation (this was discussed in detail in section 3.6.1). The excitation voltage signal prompts a response current, $\Delta i_{is}(t)$ from the solar module; as discussed in the previous chapter, impedance spectroscopy uses a voltage excitation signal and a response current signal (in potentiostatic excitation) to determine the AC impedance of a solar module, $\Delta z_{is}(t)$. These 2 signals are then processed through a Fast Fourier Transform algorithm (FFT). A Nonlinear Least Square based algorithm is then used to determine the impedance parameters for the module.
- Important factors to note:
 - With induced excitation, the BIS modulating duty cycle, $\Delta d_{is}(t)$ is used to induce the BIS voltage excitation signal, $\Delta v_{is}(t)$ at the output of the solar PV module.
 - The BIS voltage excitation signal, $\Delta v_{is}(t)$, then induces a response current $\Delta i_{is}(t)$ from the PV module.
 - The signals $\Delta v_{is}(t)$ and $\Delta i_{is}(t)$ are then used to determine the PV module's impedance $\Delta z_{is}(t)$ for impedance diagnostics.
 - Impedance measurements are only taken once power regulation is at steady state. In this context, where the instantaneous values of the output voltage of the SMC are within 10% of its average value.
 - When implementing BIS using externally generated excitation signal, the inductor current ripple is filtered out because it represents EMI noise and is undesirable; however, as will be discussed in Chapter 9, when performing EIS through the inductor current ripple as the excitation signal, the signal is not filtered out as it is desired.
 - For the solar module used, linearity is defined as voltage excitation, $\Delta v_{is}(t)$ amplitude values that are 9% – 13% of its DC PV output voltage.
 - In the figure, the input decoupling capacitor and output filter capacitors are represented as part of the SMC.

4.3 Two-level boost SMC operational analysis

As the excitation source of choice for implementing impedance spectroscopy, the two-level boost SMCs' operation was analysed. To perform impedance measurements using the converter, a small AC excitation signal will be added to the duty cycle modulating signal, thus inducing a response from the solar module which is measured to determine the impedance of the device. To ensure linearity in the implementation, both single and two-phase (multiphase) converters were analysed in continuous inductor current mode (CICM). A load which is a Li-Ion battery was used in the analysis of the converter, to represent a residential solar system with a storage option.

The single-phase converter was analysed to implement EIS on the solar module using the inductor current ripple (to be analysed in Chapter 9), while the multiphase converter was analysed to implement BIS on the module using an externally generated excitation signal, as will be analysed in Chapters 6 to 8. This is because when using an externally generated excitation signal, the inductor current ripple is not desired. However, when using the inductor current ripple to implement EIS, this is a desired parameter.

4.3.1 Single phase boost SMC

As discussed in the introduction, the boost converter is a nonlinear switch-mode DC-DC step-up converter. As illustrated by Figure 4.2, the converter will be analysed as a single input multiple output system for the purposes of implementing impedance spectroscopy. This configuration consists of 1 inductor, capacitor, switch, and diode and is powered by a solar module; because of this, a decoupling bulk capacitor is included at the input for input voltage stability. It forms the basis of a family of step-up-based converters such as, three and four-level boost converters, cascade, and quadratic boost converters, among others. The figure also shows how the EIS signal, Δd_{is} , will be superimposed or added to the DC duty cycle D_{is} to perform impedance measurements.

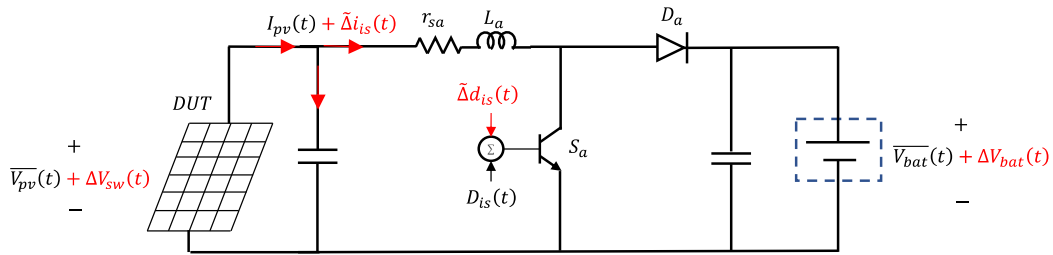


Figure 4.2 Boost switch-mode DC-DC converter topology

The converter is a nonlinear second order low pass LC filter and as an excitation source, how it responds to frequency is crucial for accurate impedance measurements. The converter has two modes operation, ON-state, and OFF-state. Figure 4.3 (a) shows that during the ON-state, the switch, S_a is closed and the inductor, L_a charges from the solar module. In this mode, the diode, D_a is reverse biased and thus, the output capacitor, C_a discharges its stored energy through the load. During this mode, the rate at which the inductor current changes is proportional to the quotient between the average DC voltage

of the module and size of the inductor. To illustrate this mathematically, consider that in this state, the PV voltage, V_{pv} is equal to voltage across the inductor, v_{L_a} , and as such:

$$\frac{dI_{L_a}}{dt} = \frac{\overline{V_{pv}}}{L_a} \quad (4.1)$$

If we now, consider the change in the inductor current during each state, (4.1) becomes:

$$\frac{\Delta I_{L_a}}{\Delta t} = \frac{\Delta I_{L_a}}{D_{dc} T_s} = \frac{\overline{V_{pv}}}{L_a} \quad (4.2)$$

Where, D_{dc} is DC time-averaged modulating duty cycle. From (4.2) the inductor current ripple can be characterized as:

$$\Delta I_{L_a} = \frac{D_{dc} \overline{V_{pv}}}{L_a f_s} \quad (4.3)$$

Where, f_s is the switching frequency resulting in, T_s . Equation (4.3) represents the typical ideal representation of the inductor current ripple as the system frequency, duty ratio and input voltage vary.

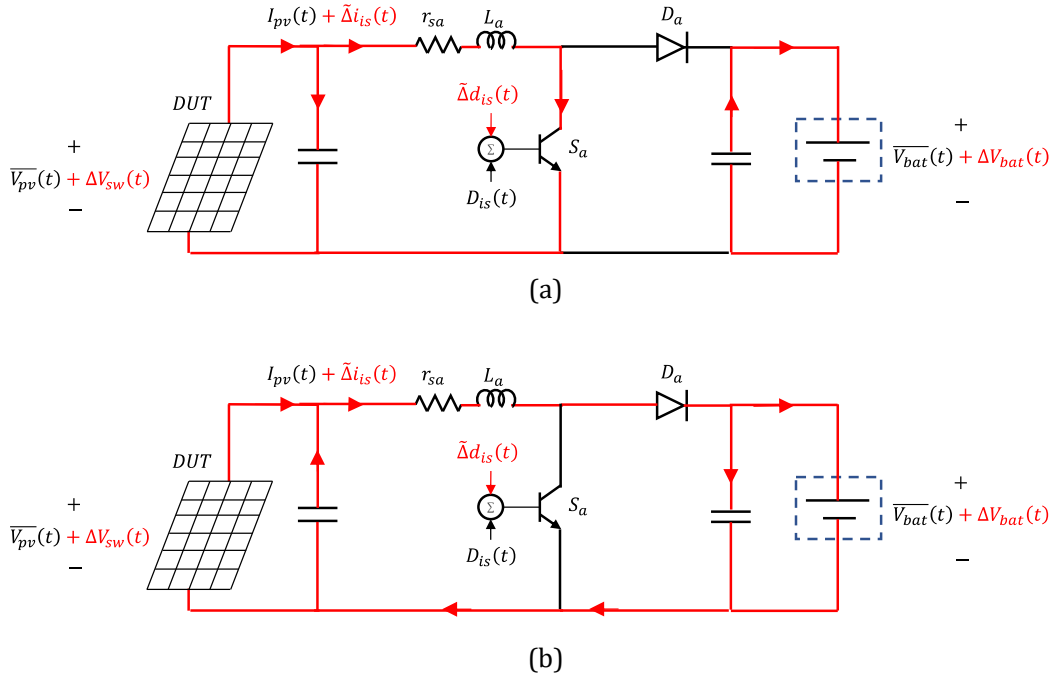


Figure 4.3 Boost SMC CICM modes of operation (a) ON-state (b) OFF-state

During the OFF state, the switch, S_a is open and the inductor's stored energy plus the output voltage of the solar module, is discharged to the load as illustrated by Figure 4.3 (b). The rate at which the inductor current drops is now proportional to the quotient of the difference between the PV voltage and battery charging voltage and the inductor. The diode is now forward biased and allows this energy to transfer to the load. The capacitor now charges up again and conducts most of the ripple created by the switch, effectively reducing the ripple at the output.

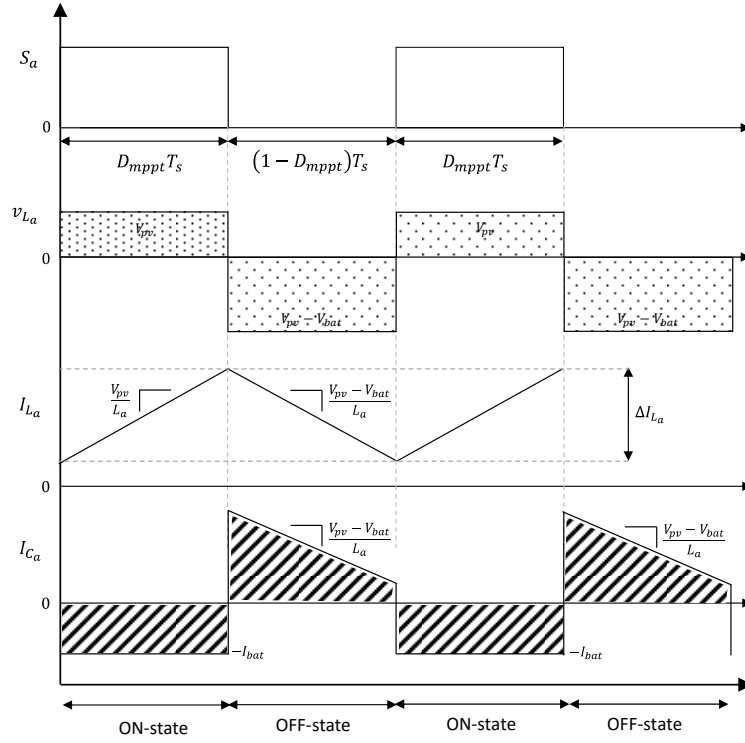


Figure 4.4 Single-phase boost SMC switch, inductor, and capacitor current signals at 50% duty cycle using PWM

Again, this can be illustrated mathematically:

$$\frac{dI_{L_a}}{dt} = \frac{\overline{V_{pv}} - \overline{V_{bat}}}{L_a} \quad (4.4)$$

If we now, consider the change in the inductor current during each state, (4.4) becomes:

$$\frac{\Delta I_{L_a}}{\Delta t} = \frac{\Delta I_{L_a}}{(1 - D_{dc})T_s} = \frac{\overline{V_{pv}} - \overline{V_{bat}}}{L_a} \quad (4.5)$$

From (4.5) the inductor current ripple can be characterized as:

$$\Delta I_{L_a} = \frac{(1 - D_{dc})(\overline{V_{pv}} - \overline{V_{bat}})}{L_a f_s} \quad (4.6)$$

This, again, shows the relation of the inductor current ripple to its parameters. The shaded area under the curve of the output capacitor current in Figure 4.4 shows the approximate total charge held by the capacitor during each state. From the physics of the capacitor, we know that the net charge, ΔQ across the capacitor during a single period must be zero i.e., the charge during each state is equal. As such, the area under the curve can be computed as:

$$\Delta Q = I_o(D_{dc}T_s) \quad (4.7)$$

Assuming a battery impedance, z_{bat} and regulated charging voltage, (4.7) becomes:

$$C_a \Delta V_{bat} = \frac{\overline{V_{bat}}}{z_{bat}} (D_{dc} T_s) \quad (4.8)$$

Where, ΔV_{bat} is the AC ripple voltage across the battery. Evaluating (4.8), the ratio of the ripple output voltage to its average value was determined to be:

$$\frac{\Delta V_{bat}}{\overline{V_{bat}}} = \frac{D_{dc}}{z_{bat} C_a f_s} \quad (4.9)$$

This ratio is important as it indicates how well the capacitor is designed for DC output voltage regulation.

4.3.2 Multiphase boost SMC analysis

In this study a 2-phase multiphase configuration, as illustrated by Figure 4.5, was analysed as the second excitation source. The converter was analysed as a ripple reduction alternative to the conventional single-phase boost converter – because as discussed in the problem statement, a lower current ripple is preferable in BIS measurements to improve the SNR and accuracy. It is worth noting that the single-phase mathematical analysis applies to the multiphase configuration of a per phase basis.

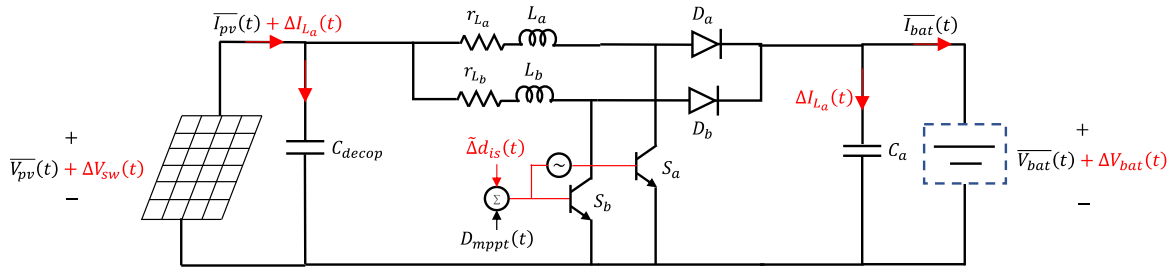


Figure 4.5 Multiphase boost switch-mode DC-DC converter topology

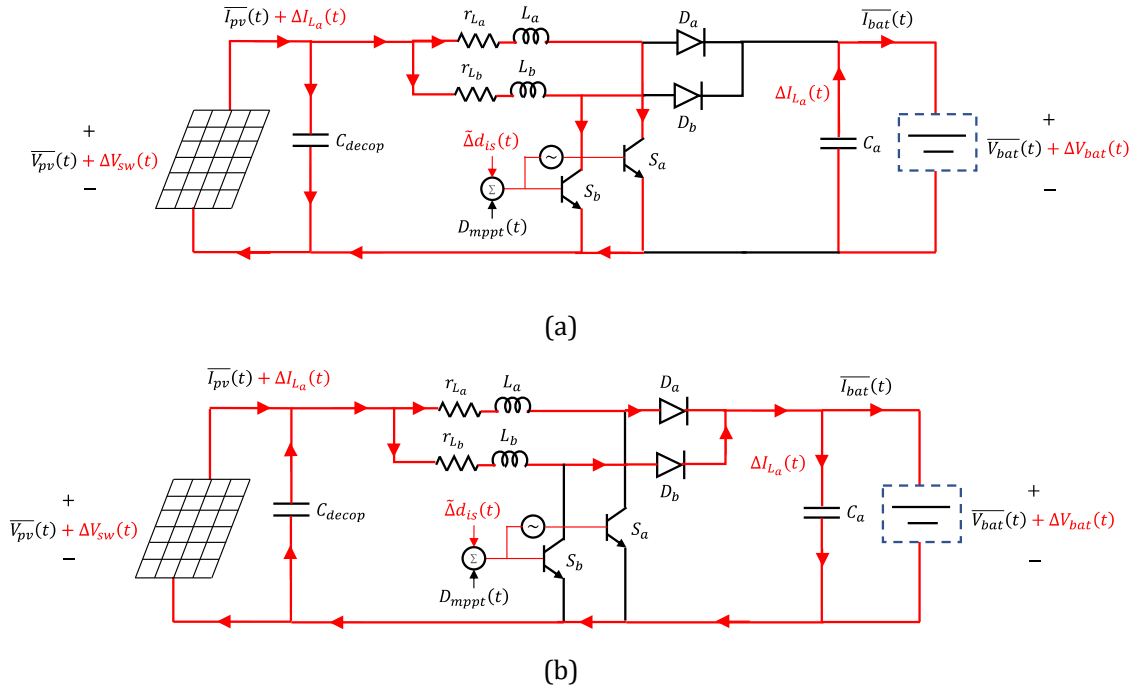


Figure 4.6 Multiphase boost SMC operational analysis (a) Mode 1 (b) Mode 2

Although this configuration now comprises of 3 storage elements, it still exhibits second order response characteristics because of the pole location of the 3rd storage element. This is desirable as it does not affect the impedance measurements and control systems. It still only has two modes, the modes 1 on 2. During the mode 1, the switch, S_a is closed and switch, S_b is open; the inductor, L_a thus, charges from the PV module, while the inductor L_b discharges through the load, since the diode, D_b is forward biased, as shown by Figure 4.6 (a). The output capacitor discharges through the load, since the diode, D_a is reverse biased. During mode 2, the switch, S_a open while switch, S_b is closed. The inductor, L_a now discharges to the load, while inductor, L_b charges from the PV module as illustrated by Figure 4.6 (b). This switching pattern results are shown in Figure 4.7 to illustrate the SMC's transient behaviour between each state.

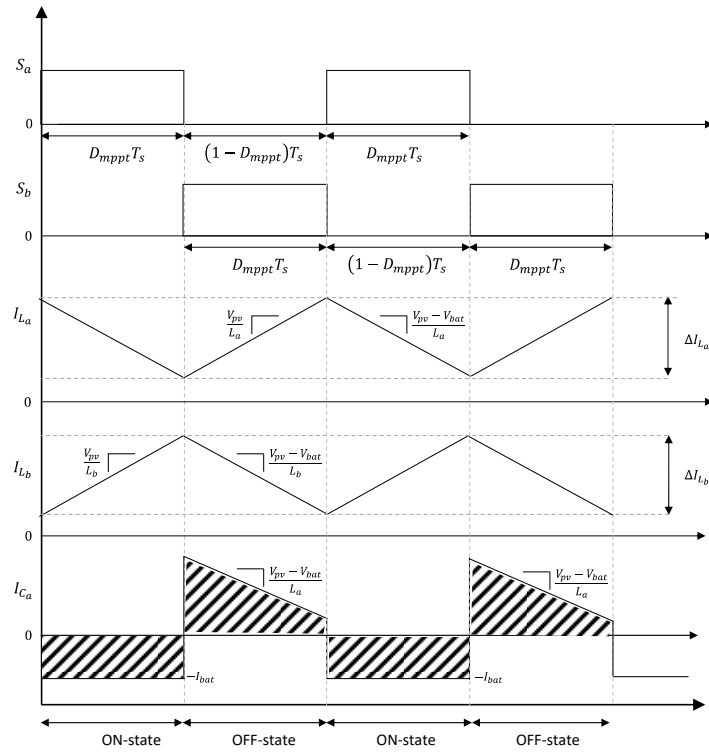


Figure 4.7 Two-phase multiphase boost SMC switches, inductors, and capacitor current signals at 50% duty cycle using PWM

4.4 Excitation source boundary conditions for linearity

To perform impedance spectroscopy on the module, the SMC, as the excitation source, must operate in CICM. This ensures that the excitation source meets the linearity requirements of impedance spectroscopy. To assess the boundary conditions, Figure 4.7 was studied. At the boundary between continuous and discontinuous mode, the average inductor current is given as:

$$\Delta I_{L_a} = 2\overline{I_{bat}} \quad (4.10)$$

Where, $\overline{I_{L_a}}$ is the average input inductor current. Because of the capacitor charge balance, the input DC current is the average of the input inductor current. From (4.3):

$$\Delta I_{L_a} = \frac{D_{dc}\overline{V_{pv}}}{L_a f_s} = 2\overline{I_{bat}} \quad (4.11)$$

The module's output current, assuming negligible inductor parasitic resistance, can be evaluated using ideal voltage transfer ratio as:

$$\overline{I_{L_a}} = \Delta I_{L_a, pk-pk} \left(\frac{D_{dc}}{1 - D_{dc}} \right) \quad (4.12)$$

And the battery charging current:

$$\overline{I_{bat}} = \frac{\overline{V_{bat}}}{z_{bat}} \quad (4.13)$$

Where R_L is the AC load resistance? Therefore from (4.10) – (4.13):

$$\frac{D_{dc}\overline{V_{pv}}}{L_a f_s} = 2 \left(\frac{D_{dc}}{1 - D_{dc}} \right)^2 \left(\frac{\overline{V_{bat}}}{z_{bat}} \right) \quad (4.14)$$

Solving for the inductor:

$$L_{a,b,min} = \frac{\overline{V_{bat}} D_{dc,min} (1 - D_{mppt,min})^2}{2 f_s \overline{I_{bat}}} \quad (4.15)$$

Where, $D_{dc,min}$ is the minimum operating duty cycle required to keep the converter in CICM. This formulation applies to both inductors in the multiphase converter. From the equation (4.15), decreasing the value of the inductor brings the converter closer to discontinuity because the inductor current decreases. Decreasing the frequency also leads to discontinuity of the converter and the lower the load current, the closer the inverter gets to discontinuity. With impedance spectroscopy now introduced, it is important to note that module's output current will change, and hence the minimum inductor current.

4.5 Two-level boost SMC power transfer analysis

The power transfer of both single and multiple-phase boost SMCs is the same, as such, this analysis applies to both configurations. Inductors in practice have DC resistance, r_{L_a} which is a parasitic resistance. This means that it is ideally desired to be zero but isn't because it is a characteristic of a real inductor. To perform this analysis, a couple of assumptions were made, firstly, the switch and capacitor parasitic resistances were considered negligible compared to the inductor's resistance; secondly, the average current across both filter and decoupling capacitors is zero. Considering Figure 4.3, and using Kirchhoff voltage laws, the clockwise voltage loop can be defined as:

$$\overline{V_{pv}} = \overline{v_{L_a}} + \overline{V_{s_a}} \quad (4.16)$$

Where, $\overline{v_{L_a}}$ is the average voltage across the inductor; and $\overline{V_{s_a}}$ is the average voltage across the switch. The average voltage appearing across the switch, can thus be described as:

$$\overline{V_{s_a}} = \frac{1}{T_s} \left(\int_0^{D_{dc}T_s} (V_{pv} - V_{pv})dt + \int_{(1-D_{dc})T_s}^{D_{dc}T_s} (V_{pv} - (V_{pv} - \overline{V_{bat}}))dt \right) \quad (4.17)$$

Equation (4.17) evaluates to an average value of:

$$\overline{V_{s_a}} = (1 - D_{dc})\overline{V_{bat}} \quad (4.18)$$

The average voltage appearing across the inductor can also be expressed as shown in (4.19).

$$\overline{v_{L_a}} = \overline{I_{L_a}} r_{L_a} \quad (4.19)$$

From Figure 4.3, the average inductor current can be expressed as:

$$\overline{I_{L_a}} = D_{mptt} \overline{I_{bat}} \quad (4.20)$$

From (4.18) – (4.20), the voltage loop equation (4.16) evaluates to:

$$\overline{V_{pv}} = D_{dc} \overline{I_{bat}} r_{L_a} + (1 - D_{dc}) \overline{V_{bat}} \quad (4.21)$$

Considering the impedance of load battery, (4.21) further evaluates to:

$$\overline{V_{pv}} = \frac{D_{dc} \overline{V_{bat}} r_{L_a}}{z_{bat}} + (1 - D_{dc}) \overline{V_{bat}} \quad (4.22)$$

Simplifying (4.22), the voltage transfer ratio of the both the boost and multiphase converters can be approximated to:

$$\frac{\overline{V_{bat}}}{\overline{V_{pv}}} = \frac{(1 - D_{dc})^{-1}}{\left(\frac{r_{L_a}}{z_{bat}(1 - D_{dc})^2} \right) + 1} \quad (4.23)$$

Equation (4.23) shows that the larger inductor's parasitic resistance, the lower the voltage ratio. Most importantly, the ratio of the parasitic resistance to the load resistance is of great importance; it shows that, at frequencies where the load impedance is large, the voltage regulation will be affected negatively as this will significantly lower the voltage transfer ratio of the regulator. When this happens, the time-averaged value of the modulating duty ratio will increase, thus resulting in poor switching utilization – this will be discussed in detail in a later section. The equation also shows that if the inductor's parasitic resistance is zero, the voltage transfer ratio of the SMC evaluates to the ideal transfer ratio. Graphically, (4.23) can be illustrated as show by Figure 4.8. Figure 4.8 (a) shows that as the parasitic resistance increases, while the load impedance is kept constant, the peak transfer ratio of the system decreases and moves to the left. This means that most of the power isn't transferred to the load but rather dissipated as heat in the converter – resulting in a less efficient system. Ideally, the curve shows no convergence and achieves infinite gain as the system's time-averaged duty ratio approaches unity. Figure 4.8 (b) shows that as the load current increases, while the parasitic resistance is kept constant, the peak transfer ratio of the system decreases and moves to the right. This means that most of the power isn't transferred to the load but rather dissipated as heat in the converter – resulting in a less efficient system. Ideally, the curve shows no convergence and achieves infinite gain as the system's time-averaged duty ratio approaches unity.

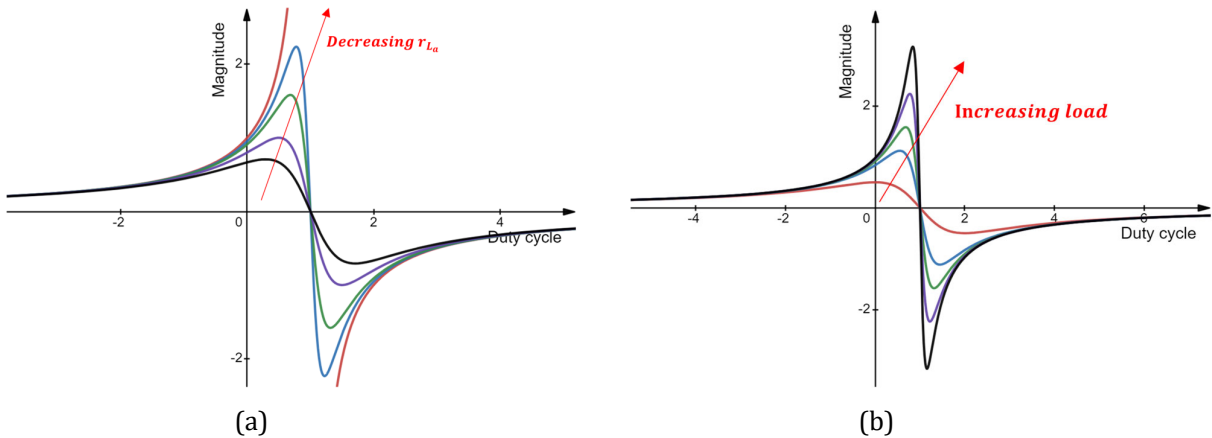


Figure 4.8 Boost SMC voltage transfer ratio (a) varying inductor parasitic resistance (b) varying load current

Figure 4.8 (b) shows that at lighter loads, the power transfer of the system is low and flat. When the load gets heavier compared to the inductor's parasitic resistance, the power transfer increases significantly. The figure shows that as the loads gets infinitely large, so does the transfer ratio; this is consistent with the theory, and simulations performed on the converter under no load conditions.

4.6 Two-level boost SMC ripple analysis

When implementing BIS on a device, it is important to keep the inductor current low. This improves the SNR of the excitation signal; thus, both configurations were analysed to quantify their inductor current ripple magnitudes. The output voltage ripple was also considered because of the importance output capacitor in design a converter with a low frequency resonant peak response.

4.6.1 Input inductor current ripple

For the single-phase boost converter, (4.3) and (4.23) were considered. Assuming negligible inductor parasitic resistance, the inductor current ripple can be re-written as:

$$\Delta I_{L_a} = \frac{D_{dc}(1 - D_{dc})\overline{V_{bat}}}{L_a f_s} \quad (4.24)$$

Rearranging (4.24) and normalizing it with respect to the peak-to-peak ripple:

$$\frac{\Delta I_{L_a} L_a f_s}{\overline{V_{bat}}} \{D_{dc}\} = D_{dc}(1 - D_{dc}) \forall D_{dc} \mid D_{dc} \in [0,1] \quad (4.25)$$

Equation (4.25) describes the inductor current ripple of the single-phase SMC as a function of its time-averaged duty ratio. It illustrates how the inductors peak-peak current ripple varies with its duty cycle. For the 2-phase multiphase converter, the same procedure was followed, and (4.3) can be rewritten as:

$$\Delta I_{L_{a,b}} = \frac{D_{dc} \left(\overline{V_{pv}} - \frac{\overline{V_{bat}}}{2} \right)}{L_a f_s} \quad (4.26)$$

Simplifying (4.26),

$$\Delta I_{L_{a,b}} = \frac{D_{dc} (2\overline{V_{pv}} - \overline{V_{bat}})}{2L_a f_s} \quad (4.27)$$

Solving (4.23) for the battery voltage and substituting the result into (4.27),

$$\Delta I_{L_{a,b}} = \frac{D_{dc} \left(2\overline{V_{pv}} - \frac{\overline{V_{pv}}}{1 - D_{dc}} \right)}{2L_a f_s} \quad (4.28)$$

Simplifying (4.28),

$$\Delta I_{L_{a,b}} = \frac{D_{dc}}{2L_f s} \left(\frac{2(1 - D_{dc})\overline{V_{pv}} - \overline{V_{pv}}}{(1 - D_{dc})} \right) \quad (4.29)$$

Finally,

$$\Delta I_{L_{a,b}} = \frac{\overline{V_{pv}}}{2L_a f_s} \left(\frac{D_{dc}(1 - 2D_{dc})}{(1 - D_{dc})} \right) \quad (4.30)$$

The normalised peak-peak ripple for the multiphase converter can be summarised as:

$$\frac{\Delta I_{L_a} L_a f_s}{\overline{V_{bat}}} \left(\frac{1}{4N} \right) \{D_{dc}\} = \begin{cases} D_{dc}(1 - 2D_{dc}), & 0 \leq D_{dc} \leq \frac{1}{N} \\ (2D_{dc} - 1)(1 - D_{dc}), & \frac{1}{N} < D_{dc} \leq 1 \end{cases} \quad (4.31)$$

Where, N is the number of phases in the SMC – two in this case. Again, equation (4.31) describes the peak-peak inductor current ripple of the multiphase converter as a function of its duty cycle. This function can be graphically represented, as shown by Figure 4.9, to better illustrate the relationship between the current ripple and SMC's operating point. The current ripple of the single-phase converter peaks at 50% duty cycle. This is not the most ideal characteristic, as the efficiency of the converter tends to be better at this duty cycle. In contrast, the current ripple of the multiphase converter convert is lowest at 50% duty cycle, but peaks at 25% and again at 75%. These peaks are 50% lower than the peak of the single-phase converter. In practical systems, duty cycles lower than 20% and higher then 80% suffer significantly from parasitic losses of the inductor, capacitor, diode, and switches. At these duty ratios, the single-phase converter has a ripple current that is 36% of its peak – this is higher than the peak of the multiphase converter. In fact, at these duty cycles, the ripple of the multiphase converter is 52% lower than the peak of the single-phase converter. This shows that in applications where the input current ripple is required to be minimal, interleaving the converter may be the most optimal solution.

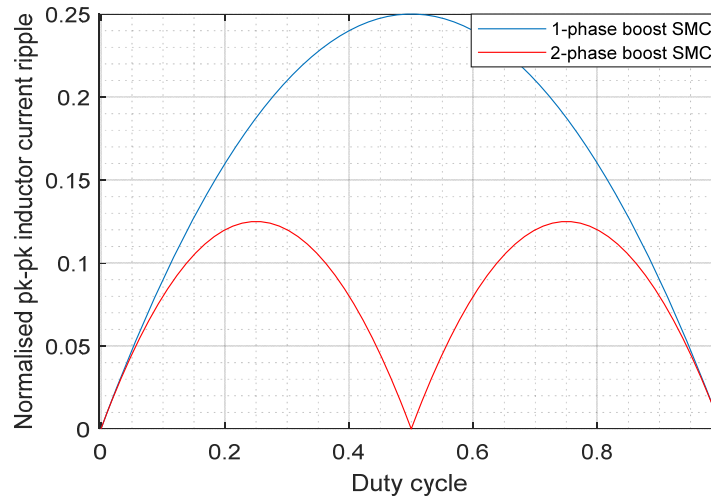


Figure 4.9 Peak to peak input current ripple comparison between the single and double phase

4.6.2 Output capacitor voltage ripple

For the single-phase converter, (4.9) was rearranged to obtain:

$$\Delta V_{bat} = \frac{D_{dc} \overline{V_{bat}}}{z_{bat} C_a f_s} \quad (4.32)$$

From (4.32) the output voltage ripple in the range of the operating points is given by:

$$\frac{\Delta V_{bat} z_{bat} C_a f_s}{\overline{V_{bat}}} \{D_{dc}\} = D_{dc} \forall D_{dc} \mid D_{dc} \in [0,1] \quad (4.33)$$

Describing the output voltage ripple as a function of the time-averaged duty cycle of the converter. The multiphase converter's output voltage ripple was resolved to be:

$$\frac{\Delta V_{bat} z_{bat} C_a f_s}{\overline{V_{bat}}} \left(\frac{1}{4N} \right) \{D_{dc}\} = \begin{cases} D_{dc}(1 - 2D_{dc}), & 0 \leq D_{dc} \leq \frac{1}{N} \\ (2D_{dc} - 1), & \frac{1}{N} < D_{dc} \leq 1 \end{cases} \quad (4.34)$$

Equation (4.34) describes the peak-peak output voltage ripple as a function of the duty cycle.

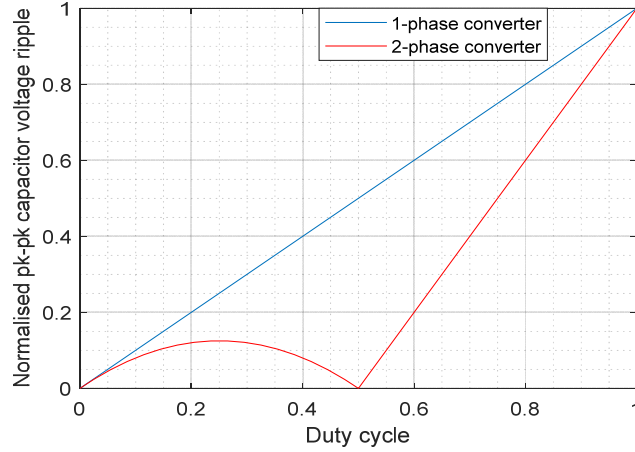


Figure 4.10 Peak to peak output voltage ripple comparison between the single and double phase

Figure 4.10 shows that for a single-phase converter, the output voltage ripple is directly proportional to the duty ratio, hence, it peaks at the theoretical maximum duty cycle. For the multiphase converter, however, the output voltage ripple peaks at 25% duty cycle before the system operating point (50% duty cycle). After this point, the rate at which this ripple changes with respect to the duty cycle is higher than that of the single-phase converter. This is because of the overlap in the switching sequences of 2-phase converter operating at higher duty cycles i.e., this configuration is specifically optimised to work at 50% duty cycle or less.

4.7 Two-level boost SMC active switch utilization analysis

As discussed earlier, solar monocrystalline modules have a wider impedance bandwidth compared to batteries; this means that the SMC must switch at frequencies greater than 100kHz . This can have a negative impact on the SMC's switching utilization as the voltage or current stress on the switching semiconductor increase. The equations discussed will be used to analyse the voltage and current stress of experienced by the chosen switch during experimental verification, to ensure that the limits of the switch are not exceeded while regulating solar power to the battery and performing impedance measurements. These stresses were analysed both configurations.

4.7.1 Wide bandgap semiconductors – a high-level review

Conventional semiconductor switches, used in almost all power converters, are made from Silicon (Si), because Si is readily available, and the assembly of the semiconductor is relatively simple. Si semiconductors, however, have the following drawbacks: lower breakdown fields – resulting in high on-state resistances and higher losses; low switching transients – limiting the ability to switch at high frequency; and low thermal conductivity – making it limited to only low current applications [7]. For these reasons, wide bandgap semiconductors were developed.

Wide bandgap (WBG) semiconductor devices have a bandgap that is wider than traditional semiconductors. The bandgap is the difference in energy between a material's conduction and valence bands [8]. In for a single electron to conduct, it must have sufficient energy to transition between the two bands. Materials that conduct electricity either have overlapping bands or narrow bandgaps. Although this is ideal, the bandgap size decreases with increasing energy, which affects the material's functionality; thus, wide bandgap semiconductors have a bandgap that is wide enough such that material can conduct electricity while having the capacity to do so at high temperatures [9]. This ability ensures that the devices can achieve better efficiencies, operate at higher switching speeds, handle higher operating voltages and temperatures, and have a smaller footprint [10]. The most common WBG semiconductors are listed, as shown in Table 4.1 [11].

Table 4.1 Comparison of conventional and wide bandgap semiconductor properties

Property	Si	GaAs	6H-SiC	4H-SiC	GaN	Diamond	AlN
Bandgap Energy (eV)	1.12	1.42	3.00	3.36	3.44	5.45	6.20
Electric breakdown field (MV/cm)	0.30	0.40	2.50	2.00	3.80	10.00	12.00
Electron mobility (cm^2/Vs)	1500	8500	420.0	950.0	1000	3800	300.0
Saturated electron drift velocity ($\times 10^7$)	1.00	1.20	2.00	2.00	2.50	2.70	1.70
Dielectric constant	11.2	13.1	9.70	10.0	9.50	5.50	8.50
Thermal conductivity (W/cmK)	1.50	0.46	4.90	4.90	1.30	22.0	2.85

The material properties are obtained from experimental methods and as such, they differ between sources; however, but it is evident that WBG semiconductors have significant advantages over traditional narrow bandwidth semiconductors. WBG semiconductors have a higher breakdown field which

allows them to have thin drift regions compared to traditional semiconductors, thus resulting in low on-resistances; this also improves the drain current capability of the semiconductor, which results in low input and output application capacitances. The devices also have high saturation velocities and low parasitic capacitances, which allows for faster switching transients – as desired in this application. SiC and GaN semiconductors are among the most common WBG devices. Based on the table SiC WBGs have higher thermal conductivity means the device will transfer heat rapidly to a heatsink for better device cooling, compared to GaN WBGs. Given that all other parameters are similar between the 2 types of WBGs, the superior thermal conductivity of the SiC makes it ideal for high switching applications are high currents, and hence, this application.

4.7.2 Voltage stress

The voltage stress on the single-phase converter can be computed by considering (4.18),

$$V_{mos} = \overline{V_{bat}}(1 - D_{dc})(1 + \Delta V_{bat\%}) \quad (4.35)$$

Where, $\Delta V_{bat\%}$ is the ratio of the output voltage ripple to its average value. Substituting (4.32) into (4.35):

$$V_{mos} = \overline{V_{bat}}(1 - D_{dc}) \left(1 + \left(\frac{D_{dc} \overline{V_{bat}}}{2(z_{bat} \overline{V_{pv}} C_a f_s)} \right) \right) \quad (4.36)$$

This equation is important because it shows the switching frequency as a parameter of the voltage stress across the switch, which affirms the hypothesis made about problems with switching at high frequencies.

4.7.3 Current stress

To determine this stress, the maximum current that can be exerted on to the switch was determined. The duty cycle at which the maximum current will occur is:

$$I_{mos} = \overline{I_{bat}}(1 + \Delta I_{sw\%}) \quad (4.37)$$

Where, $\Delta I_{sw\%}$ is the ratio of the inductor current ripple to its average value. The maximum current for the switch is thus:

$$I_{mos} = \overline{I_{bat}} \left(\frac{1}{I_{La}} \left(1 + \frac{(1 - D_{dc})(\overline{V_{pv}} - \overline{V_{bat}})}{L_a f_s} \right) \right) \quad (4.38)$$

This is the maximum current that will be experienced by the switch.

4.7.4 Switch utilization

The switching utilization is given by:

$$\delta_{mos} = \overline{I_{bat}V_{bat}} \left(\sum_{i=1}^N V_{mos,i} I_{mos,i} \right)^{-1} \quad (4.39)$$

Since the converter only has 1 switch, the utilization can thus be evaluated to:

$$\delta_{mos} = \frac{(1 - D_{dc})^{-1}}{\left(1 + \left(\frac{D_{dc} \overline{V_{bat}}}{2(z_{bat} \overline{V_{pv}} C_{afs})} \right) \right) \left(\frac{1}{I_{La}} \left(1 + \frac{(1 - D_{dc})(\overline{V_{pv}} - \overline{V_{bat}})}{L_a f_s} \right) \right)} \quad (4.40)$$

Which is the switch utilization for the converter, which can be evaluated for the value at the module's bandwidth. For the multiphase converter, the switching utilization is dependent on the number of phases it has. i.e.

$$\delta_{mos} = \frac{(1 - D_{dc})^{-1}}{N \left(1 + \left(\frac{D_{dc} \overline{V_{bat}}}{2(z_{bat} \overline{V_{pv}} C_{afs})} \right) \right) \left(\frac{1}{I_{La}} \left(1 + \frac{(1 - D_{dc})(\overline{V_{pv}} - \overline{V_{bat}})}{L_a f_s} \right) \right)} \quad (4.41)$$

For each switching. Hence the total switching utilization is linearly related to the number of phases such that, the larger the number of phases, the lower the stress on each switch. This gives the possibility of using lower powered semiconductor switches in the devices which may reduce the size and cost of the system.

4.8 Boost SMC power regulation and perturbation signal preliminary design

The preliminary design of both single-phase and multiphase SMCs was detailed through specifications obtained from a residential hybrid solar system with a storage option. This design is preliminary because it is based on the simulation environment assumptions; during experimental verification, the final design, accounting for practical consideration, and component selection will be evaluated.

4.8.1 Power regulation design specifications

Impedance spectroscopy is to be implemented while the solar module delivers power to a battery pack through the SMC, as such the specifications for the SMC required to regulate the solar power delivered to the load battery are detailed in Table 4.2. A 12V load Li-Ion battery with a maximum charging current of 2.68A, a maximum charging voltage of 14.5V and minimum charging voltage of 13.2V, was chosen for simulation purposes – this emulated the battery to be used in the experimental verification.

Table 4.2 Proposed system specifications

Specification type	Specification
System power	80W
Input voltage range	10 – 21V _{DC}
Carrier switching frequency	500kHz
Rated module current	5.35
Max DC load/charging current	2.68A
Inductor DCR r_L	100mΩ
Capacitor ESR r_C	30mΩ
Output voltage ripple	$\leq 2\%$ of V_o
Inductor ripple Δi_L	30% of $i_{L,max}$
Maximum duty cycle	83.69%
Nominal duty cycle	50%
Minimum duty cycle	16.30%

The inductor DCR and capacitor ESR values were chosen for design and simulation purposes. They were chosen based on general values obtained for these devices; during experimental verification, the actual resistances will be measured and used for the design. The duty cycle maximum percentage was chosen based on the nominal input voltage from the solar module and the maximum charging voltage for a 12V load battery; while the minimum duty cycle was chosen based on the nominal input solar voltage from the module and minimum voltage required to charge a 12V battery. The percentage of the inductor current ripple was chosen to allow for maximal power transfer without increasing the size of the inductor significantly. The output voltage ripple was chosen to low to ensure a larger DC component for effective solar power delivery. The specification in Table 4.2 were used to design the system such that the solar module delivers and stores power in a battery storage system. The nominal duty cycle was chosen such that a 36-cell module could charge a 12V battery.

4.8.2 Solar module selection

A shell PowerMax® Ultra 80-P monocrystalline solar module was used as the DUT. The module is an 80W_p source with an open circuit voltage of 21.8V and short circuit current of 5.35A. It's rated at 16.9V and 4.76A at its maximum power. Its simulated characteristic curves are as illustrated. Due to the limitation of the MATLAB environment, only the temperature variation is shown, however, this does not affect the simulation results or specifications. Figure 4.11 shows that as the temperature varies, the module's open circuit voltage varies significantly; although not at the same rate of change, the module's current at maximum power varies as well, as illustrated by Figure 4.12. Although there is change in the module's short circuit current, it is minimal, with a 0.5% change from STC conditions, compared to a 9% change in the module's short circuit current.

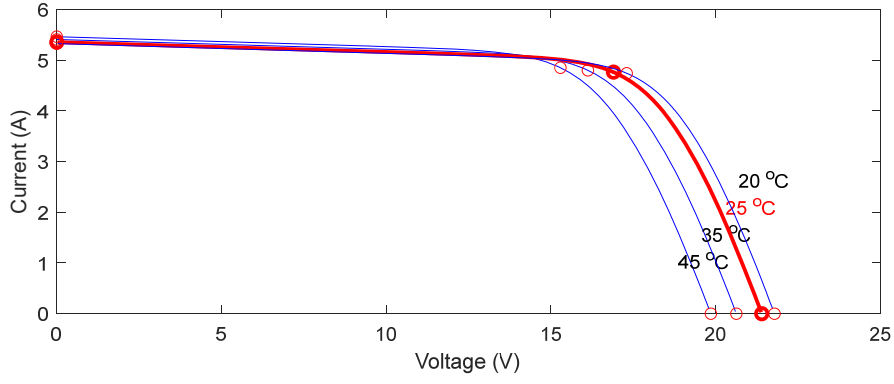


Figure 4.11 Shell PowerMax I/V characteristics

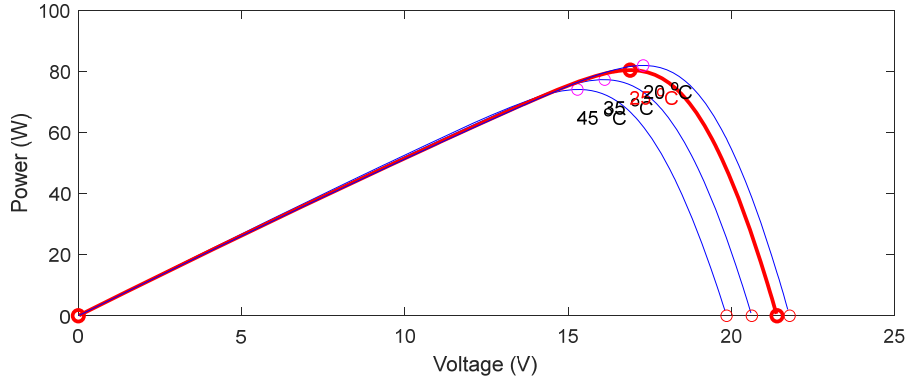


Figure 4.12 Shell PowerMax P/V characteristics

4.8.3 Boost SMC power regulation design

i. Inductor selection

The inductance was designed rearranging (4.3) and using specifications from Table 4.2. The rearranged equation becomes:

$$L_a = \frac{D_{dc} \overline{V_{pv}}}{\Delta I_{L_a} f_s}$$

The nominal peak-peak ripple percentage, as indicated in Table 4.2, was calculated considering that a good estimation for the peak-peak ripple current is between 10% and 40% of the DC average value, as illustrated by (4.42), while ensuring that the ripple current does not exceed 50% of the module's output current at full load, ensures stability in the control system of the converter.

$$\Delta I_{sw} = \Delta I_{sw\%} \left(\frac{I_{bat,max} V_{bat}}{\overline{V_{pv}}} \right) \quad (4.42)$$

Moreover, the inductor size is designed such that the input current is continuous for load currents above 5% – 15% of full load current. The minimum current required for continuous mode also needs to be considered since impedance measurements must be taken in continuous mode to meet LTI

constraints. It is worth noting that a higher ripple current increases the inductor proximity and core losses. However, a higher value of the inductor, with a low ripple, increases the DC resistance losses. Consideration a converter switching a nominal switching speed of, $500kHz$ the inductance value was designed to be $26\mu H$.

ii. **Output capacitor selection**

The output capacitor was designed to filter out AC components of both the induced small signal output voltage, because of the injection of the BIS signal, and the high frequency switching signals. The best practice is to use capacitors with low DC resistance to better minimize the output voltage ripple due to the switching signal. The voltage ripple introduced by the DC resistance can be characterized as:

$$\Delta V_{bat} = r_{c,o} \left(\frac{\overline{I_{bat,max}}}{1 - D_{dc}} + \frac{\Delta I_{sw}}{2} \right) \quad (4.43)$$

Where, $r_{c,o}$ is the DC resistance. A low DC resistance thus, minimizes this ripple. Equation (4.9) was rearranged such that:

$$C_a = \frac{\overline{V_{bat}} D_{dc}}{z_{bat} \Delta V_{bat} f_s}$$

and using Table 4.2 the value of this capacitance was evaluated. The optimal capacitance value was chosen to be $349\mu F$.

iii. **Input capacitor selection**

Because of the BIS signal superposition, there is a small signal AC instantaneous power drawn from the PV module, as well as the AC power from the switching transients of the converter. As such, the total power drawn from the module can be characterized as a summation of the DC power, the BIS pulsating power, and the switching ripple power, as illustrated by (4.44):

$$P_{pv}(t) = \Sigma \left\{ \begin{array}{l} \overline{V_{pv}(t)} \overline{I_{pv}(t)} \\ \frac{\langle \tilde{\Delta} v_{in} \rangle \langle \tilde{\Delta} i_{in} \rangle}{2} (\cos \beta_{bis} - \cos(2\omega_{bis} t + \beta_{bis})) \\ \frac{D_{dc} (\langle \tilde{\Delta} v_{sw} \rangle \langle \tilde{\Delta} i_{sw} \rangle)}{2} (\cos \beta_{sw} - \cos(2\omega_{sw} t + \beta_{sw})) \end{array} \right\} \quad (4.44)$$

Where, $\langle \tilde{\Delta} v_{in} \rangle$ is the excitation signal peak magnitude; $\langle \tilde{\Delta} i_{in} \rangle$ is the response signal peak magnitude; $\langle \tilde{\Delta} v_{sw} \rangle$ is the peak voltage ripple due to the switching states and the inductor; $\langle \tilde{\Delta} i_{sw} \rangle$ is the peak inductor current ripple; β_{bis} is the phase shift between the excitation signal and response signal – assuming the excitation signal as a reference; and β_{sw} is the phase shift between the switching ripple voltage and current. These two AC power components are dissipated through the decoupling capacitor and thus reduce

the efficiency of the system. The peak inductor current ripple was designed to be 20% of the battery's maximum charging current, at a voltage transfer ratio of 2 – the system's DC operating point. At this operating point, the 2-phase multiphase converter reduces this ripple by half. The converter's switching frequency is 6 times the maximum impedance frequency, thus ensuring a smaller resulting ripple because of the inverse proportionality relationship. This further reduces the contribution of the switching ripple power – and hence, its contribution to the energy stored in the decoupling capacitor. Thus, to ensure power balance in the system, the decoupling capacitor was designed based on the DC and BIS power. It was designed such that it was equivalent to the energy that can be stored by the device as shown by (4.45).

$$C_{decop} = \frac{\overline{P_{pv}}}{4\pi\overline{V_{pv}} \left(\max_{f_{is}}(f_{is} + f_{sf}) \right) \left(\max_{v_{in}}(\tilde{\Delta}v_{in}) \right)} \quad (4.45)$$

Where, f_{sf} is a safety design factor – which is 5% of the maximum broadband frequency. The peak voltage amplitude of the induced excitation signal was used to define the maximum ripple voltage of the decoupling capacitor. The effect of this capacitance on impedance measurements is the attenuation of higher frequency values. This is because the capacitor must be large enough for input voltage stability to the converter, but also needs to be small enough to allow a passage of high frequency impedance values. This trade-off allowed for an iterative selection of the capacitor to best satisfy both conditions – with (4.45) as the reference value. The selected value was $1.72mF$.

iv. **Per phase multiphase boost converter design**

The design of the per phase values, for the multiphase converter, follow the same approach as the single-phase converter – with a few considerations. Firstly, the per phase average inductor current is given by:

$$\Delta I_{Lp} = \frac{\overline{I_{bat}}}{p(1 - D_{dc,max})} \quad (4.46)$$

Where, p is the converter phase; and $D_{dc,max}$ is the maximum system duty cycle. While the input module current is given by:

$$\Delta I_L = \sum_{p=1}^N \frac{\overline{I_{bat}}}{p(1 - D_{max})} \quad (4.47)$$

Where, N is the number of phases in the converter.

4.9 Power regulation control – 1st tier control design

As mentioned in the introduction, the control system being undertaken comprises of 3 tiers, one of which is power regulation control subsystem. Power regulation control was required to ensure

regulated power transfer between the solar module and load, and more importantly, achieving a wide current loop bandwidth for excitation signal induction. In a solar and battery hybrid system, the two devices are connected through a power optimizer. Hence, the one of the most evident integrations of the online impedance extracted systems would be embedded with a power optimizer; thus, the system was designed to emulate such an integration.

4.9.1 Design specifications

The design specifications were tabulated as shown in Table 4.3. The steady state error, bandwidth, settling time, phase, and gain margins.

Table 4.3 Control design specifications

Specification Type	Specification
Current loop Steady State Error	$\pm 5\%$
Voltage loop Steady State Error	$\pm 10\%$
Voltage loop Percentage Overshoot	$\leq 10\%$
Current loop Percentage Overshoot	$\leq 13\%$
Voltage Loop Bandwidth	$\leq 217\text{Hz}$
Current Loop Bandwidth	$\cong 50\text{kHz}$
Current Loop Phase margin	$\geq 62^\circ$
Voltage Loop Phase margin	$\geq 65^\circ$
Gain Margin	$\geq 6\text{dB}$
Damping factor	$0.6 - 0.8$
Settling time	$< 30\text{ms}$

The specifications were determined based on the reasoning as detailed in this section. It is worth noting that some of the specifications will change during experimental verification to accommodate more realistic margins based on the limitations of the components and other system constraints.

i. Steady state error

The steady state error in the output voltage is required to vary between a maximum of $\pm 10\%$ to accommodate the variation in the output voltage due to impedance measurements. This will be discussed in more detail in Chapter 6. The steady state error in the induced input current, however, is required to vary only between $\pm 5\%$. This is due to the linearity required for impedance spectroscopy; this value was determined experimentally, and will be discussed in Chapter 5, during experimental verification.

ii. Percentage overshoot

The overshoot in the voltage was chosen such that the phase margin could be maximised, without overdamping the loop. Larger values were avoided to ensure good system damping and reduce system oscillations for a better steady state response. For the current loop, the overshoot was based on the

maximum allowable excitation amplitude for the solar module to remain within its linear range during impedance measurements.

iii. **Phase margin**

The phase margin (PM) was based on the percentage overshoot as shown by equation [12]:

$$PM = 75 - \%OS \quad (4.48)$$

Substituting for the values gave the PM shown in the equation.

iv. **Gain margin**

The gain margin was chosen based on the rule of thumb $(2 - 10)dB$ [13], for boost converters. In the experimental verification, a more thorough analysis into the choice of gain margin will be undertaken, factoring in the equipment used.

v. **Loop bandwidths**

The current loop bandwidth was chosen as such to allow for the maximum passage of impedance frequency. This will maximise the amplitude at high frequencies for impedance estimation. Although the maximum expected excitation frequency is $100kHz$ – the bandwidth of the solar module as discussed in Chapter 1, designing the loop bandwidth to achieve an equivalent bandwidth would be limited by the practical system. Hence, a good design generally chooses a bandwidth that is between, 10% and 20% of the converter's switching speed – thus, 10% of $500kHz$ was chosen as the worst-case design, as quoted in Table 4.3. The voltage loop bandwidth was chosen as low as possible to reduce the effect of the right-half-plane zero; this will be discussed in more detail during the control design section.

4.9.1 System modelling

Impedance estimation will be conducted using small ac perturbations, such that both PV module and converter can be assumed to be pseudo-linear around a desired operating point. Using the boost converter as an excitation source requires that it operates in continuous inductor current mode (CICM). This enables the response current to be measured accurately. Since the converter will be operated in CICM and is time-invariant, it can be represented by a state space model as illustrated:

$$\dot{x} = Ax + Bu \quad (4.49)$$

$$y = Cx + Du \quad (4.50)$$

Where, x is the state vector; u is the input vector; y is the output vector; and A, B, C, D are matrices that describe the behavior the PV module's boost converter dynamic system. The derivative of the state vector has the same dimensions as the number of storage elements in the converter i.e.

$$\dot{x} = \begin{bmatrix} \frac{di_{in}(t)}{dt} \\ \frac{dv_{bat}(t)}{dt} \end{bmatrix} \quad (4.51)$$

Although the boost converter is time-invariant, it is still non-linear because of its switching states. State space modelling can be used to average the converter between its two states. This can be represented as:

$$\dot{x} = A_1x + B_1u \quad (4.52)$$

during the converter's ON-time and:

$$\dot{x} = A_2x + B_2u \quad (4.53)$$

during the converter's OFF-time – averaged over the switching signal period, T_s . The average system can therefore be written as:

$$A = A_1D_{dc} + B_1(1 - D_{dc}) \quad (4.54)$$

$$A = A_2D_{dc} + B_2(1 - D_{dc}) \quad (4.55)$$

During the ON-time, the PV-converter system can be described as follows:

$$V_{dc}(t) - i_{in}(t)(\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}) - L \frac{di_{in}(t)}{dt} = 0 \quad (4.56)$$

$$C \frac{dv_{bat}(t)}{dt} + \frac{v_{bat}(t)}{\tilde{\Delta}z_{bat}} - I_{dc}(t) = 0 \quad (4.57)$$

In matrix form, (4.56) and (4.57) become:

$$\begin{bmatrix} \dot{i}_{in}(t) \\ \dot{v}_{bat}(t) \end{bmatrix} = \begin{bmatrix} -(\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}) & 0 \\ 0 & -\frac{1}{\tilde{\Delta}z_{bat}C} \end{bmatrix} \begin{bmatrix} i_L(t) \\ v_c(t) \end{bmatrix} + \begin{bmatrix} \frac{1}{L} & 0 \\ 0 & \frac{1}{C} \end{bmatrix} \begin{bmatrix} V_{dc}(t) \\ I_{dc}(t) \end{bmatrix} \quad (4.58)$$

Where, $\begin{bmatrix} -(\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}) & 0 \\ 0 & -\frac{1}{\tilde{\Delta}z_{bat}C} \end{bmatrix} = A_1$ and $\begin{bmatrix} \frac{1}{L} & 0 \\ 0 & \frac{1}{C} \end{bmatrix} = B_1$

During the OFF time:

$$V_{dc}(t) - i_L(t)(\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}) - L \frac{di_L(t)}{dt} - v_c(t) = 0 \quad (4.59)$$

$$-i_L(t) + C \frac{dv_c(t)}{dt} + \frac{v_c(t)}{\tilde{\Delta}z_{bat}} - I_{dc}(t) = 0 \quad (4.60)$$

In matrix form, (4.59) and (4.60) become:

$$\begin{bmatrix} \dot{i}_L(t) \\ \dot{v}_c(t) \end{bmatrix} = \begin{bmatrix} -(\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}) & \frac{1}{L} \\ \frac{1}{C} & -\frac{1}{\tilde{\Delta}z_{bat}C} \end{bmatrix} \begin{bmatrix} i_L(t) \\ v_c(t) \end{bmatrix} + \begin{bmatrix} \frac{1}{L} & 0 \\ 0 & \frac{1}{C} \end{bmatrix} \begin{bmatrix} V_{dc}(t) \\ I_{dc}(t) \end{bmatrix} \quad (4.61)$$

Where, $\begin{bmatrix} -(\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}) & \frac{1}{L} \\ \frac{1}{C} & -\frac{1}{\tilde{\Delta}z_{bat}C} \end{bmatrix} = A_2$ and $\begin{bmatrix} \frac{1}{L} & 0 \\ 0 & \frac{1}{C} \end{bmatrix} = B_2$

Using (4.54) and (4.55), (4.58) and (4.61) become:

$$A = \begin{bmatrix} -(\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}) & 0 \\ 0 & -\frac{1}{\tilde{\Delta}z_{bat}C} \end{bmatrix} D_{dc} + \begin{bmatrix} -(\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}) & \frac{1}{L} \\ \frac{1}{C} & -\frac{1}{\tilde{\Delta}z_{bat}C} \end{bmatrix} (1 - D_{dc}) \quad (4.62)$$

Which simplifies to:

$$A = \begin{bmatrix} -(\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}) & \frac{(1 - D_{dc})}{L} \\ \frac{(1 - D_{dc})}{C} & -\frac{1}{\tilde{\Delta}z_{bat}C} \end{bmatrix} \quad (4.63)$$

The other matrix becomes:

$$B = \begin{bmatrix} \frac{1}{L} & 0 \\ 0 & \frac{1}{C} \end{bmatrix} D_{dc} + \begin{bmatrix} \frac{1}{L} & 0 \\ 0 & \frac{1}{C} \end{bmatrix} (1 - D_{dc}) = \begin{bmatrix} \frac{1}{L} & 0 \\ 0 & \frac{1}{C} \end{bmatrix} \quad (4.64)$$

The state space model of the PV module's boost converter then becomes:

$$\begin{bmatrix} \dot{i}_{in}(t) \\ \dot{v}_{bat}(t) \end{bmatrix} = \begin{bmatrix} -(\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}) & \frac{(1 - D_{dc})}{L} \\ \frac{(1 - D_{dc})}{C} & -\frac{1}{\tilde{\Delta}z_{bat}C} \end{bmatrix} \begin{bmatrix} i_{in}(t) \\ v_{bat}(t) \end{bmatrix} + \begin{bmatrix} \frac{1}{L} & 0 \\ 0 & \frac{1}{C} \end{bmatrix} \begin{bmatrix} V_{dc}(t) \\ I_{dc}(t) \end{bmatrix} \quad (4.65)$$

Equation (4.65) is still non-linear but small signal analysis can be used to approximate a linear model around a designed operating point. The converter can be modelled using large DC and small AC signals such that:

$$\left. \begin{aligned} u &= U + \tilde{u} \\ x &= X + \tilde{x} \\ d &= D_{dc} + \tilde{d}_{is} \end{aligned} \right\} \quad (4.66)$$

Both AC and DC components are known quantities. Incorporating both large and small signals into the state equation results in the following:

$$\dot{X} + \tilde{X} = x[A_1 d + A_2(1 - d)] + [B_1(d) + B_2(1 - d)]u \quad (4.67)$$

To linearize the converter, second order terms can be assumed to be small enough i.e., the product of $\tilde{u}$, $\tilde{x}$ and $\tilde{\Delta}d_{bis}$ can be neglected. The removal of these second order terms linearizes the equation. It can also be assumed that the induced voltage excitation and current response signals are small enough such that, U is the only term considered for the input vector. Since impedance estimation is conducted while the system is in steady state, $\dot{X} = 0$. As such:

$$\tilde{X} = A\tilde{x} + \tilde{\Delta}d_{is}[(A_1 + A_2)X + (B_1 + B_2)U] \quad (4.68)$$

To obtain the system transfer function, the Laplace of the system was taken such that the state equation becomes:

$$\tilde{x}(s) = [sI - A]^{-1} \tilde{\Delta}d_{is}[(A_1 + A_2)X + (B_1 + B_2)U] \quad (4.69)$$

Substituting and simplifying, the LTI current loop dynamic system can be described as:

$$P^*_{pv}(s) = \frac{\tilde{\Delta}i_{in}(s)}{\tilde{\Delta}d_{in}(s)} = \frac{\frac{\overline{V_{bat}}}{L_a}s + \frac{V_c}{\tilde{\Delta}z_{bat}L_aC_a} + \frac{I_L(1 - D_{dc})}{L_aC_a}}{s^2 + \left(\frac{\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}}{L_a} + \frac{1}{\tilde{\Delta}z_{bat}C_a}\right)s + \left(\frac{\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}}{\tilde{\Delta}z_{bat}L_aC_a} + \frac{(1 - D_{dc})^2}{L_aC_a}\right)} \quad (4.70)$$

The dynamic system in (4.70) describes the module's change in its output current from an injected duty ratio signal – thus, implying induced current excitation. As mentioned earlier, voltage excitation is desired for the PV module, as it will be operated as a current source during impedance extraction. Impedance spectroscopy on the module is designed such that the system remains in its pseudo-linear during the process – as such, the induced excitation and response signals at the low voltage end, can be linearly related by the module's impedance.

$$\frac{\tilde{\Delta}v_{in}(s)}{\tilde{\Delta}d_{in}(s)} = \tilde{\Delta}z_{pv}(s) \quad (4.71)$$

Thus, after substituting the result into (4.70) and simplifying, the induction of the module's voltage excitation signal from an injected duty cycle perturbation signal, can be described as illustrated by:

$$P^i_{pv}(s) = \frac{\tilde{\Delta}v_{in}(s)}{\tilde{\Delta}d_{in}(s)} = \frac{\frac{\tilde{\Delta}z_{pv}\overline{V_{bat}}}{L_a} \left[s + \left[\frac{1}{\tilde{\Delta}z_{bat}C_a} + \frac{I_L(1 - D_{dc})}{C_a\overline{V_{bat}}} \right] \right]}{s^2 + \left[\frac{\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}}{L_a} + \frac{1}{\tilde{\Delta}z_{bat}C_a} \right] s + \left[\frac{\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}}{\tilde{\Delta}z_{bat}L_aC_a} + \frac{(1 - D_{dc})^2}{L_aC_a} \right]} \quad (4.72)$$

The model shows an LTI, type zero, proper and minimum phase plant – with a system gain characterized by the PV module's impedance parameters, the battery charging voltage and system input inductance.

$$P_{pv}^v(s) = \frac{\tilde{\Delta}v_{bat}(s)}{\tilde{\Delta}d_{in}(s)} = \frac{\frac{\bar{V}_{bat}(1-D_{dc})}{L_a C_a} - \frac{\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}}{L_a C_a} - \frac{I_L}{C_a} s}{s^2 + \left(\frac{\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}}{L_a} + \frac{1}{\tilde{\Delta}z_{bat} C_a} \right) s + \left(\frac{\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}}{\tilde{\Delta}z_{bat} L_a C_a} + \frac{(1-D_{dc})^2}{L_a C_a} \right)} \quad (4.73)$$

While (4.73) describes the change in the converter's output voltage from an injected duty ratio.

4.9.2 Closed loop system analysis

The plants described in (4.72) and (4.73) were used to assess the stability and response of the input current and output voltage's closed loop. The following architecture was used for the control system. The architecture was chosen because, as observed in the previous section, the injection of a single small signal duty ratio induces 2 output signals, current at the input, and voltage at the output. This analysis is performed to determine which, if any, specifications both voltage and current loops meet without the design of a controller; this will ensure that an optimal controller is designed without over-designing for a specification that the system meets without control compensation.

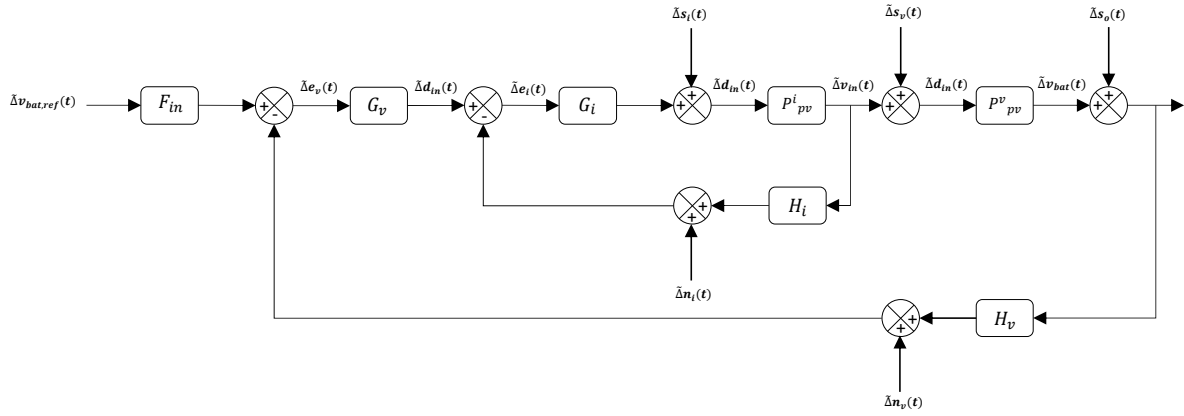


Figure 4.13 Control system closed loop architecture

From Figure 4.13, the inner current loop is responsible for the injection of the impedance spectroscopy modulating duty ratio, $d_{in}(t)$ – which can take several forms depending on the type of excitation chosen, while the outer voltage loop is responsible for ensuring regulated output voltage: both plants without compensation. The 3 plots show that the closed loop is stable. The root locus shows the region where the closed loop poles must lie to achieve the desired results. The closed loop poles start at open loop poles and terminate at open loop zeros. With a closed loop gain of 1, the closed loop poles sit out the desired region, hence a controller must be designed to ensure the system meets its specifications. The plot also shows that the system's zero is on the right half plane (RHP) near the origin – making the plant, non-minimum phase. The RHP zero phenomenon of a boost converter has been researched in

several work [14] – [16]. This occurs because, when the converter operates at high duty cycles, the inductor and load are disconnected for a longer period i.e., as reviewed earlier in the chapter, during the ON-time, the diode D_a is reverse biased and ensures that current does not flow from the inductor to the load. The output voltage thus, drops while the control action tries to increase it; this then leads to the output dip as observed in Figure 4.14 (c). The output only increases once the transients associated with the zero have elapsed; the frequency of which is: $f_z = \tilde{\Delta}z_{bat}(1 - D_{dc,max})^2 / 2\pi L_a$ – which is also the location of the zero on the root locus. This means that when the system is given a command, it will not response immediately to the control action; thus, to ensure stability in the system, the bandwidth of the voltage loop must be reduced such that it is less than the frequency of the zero.

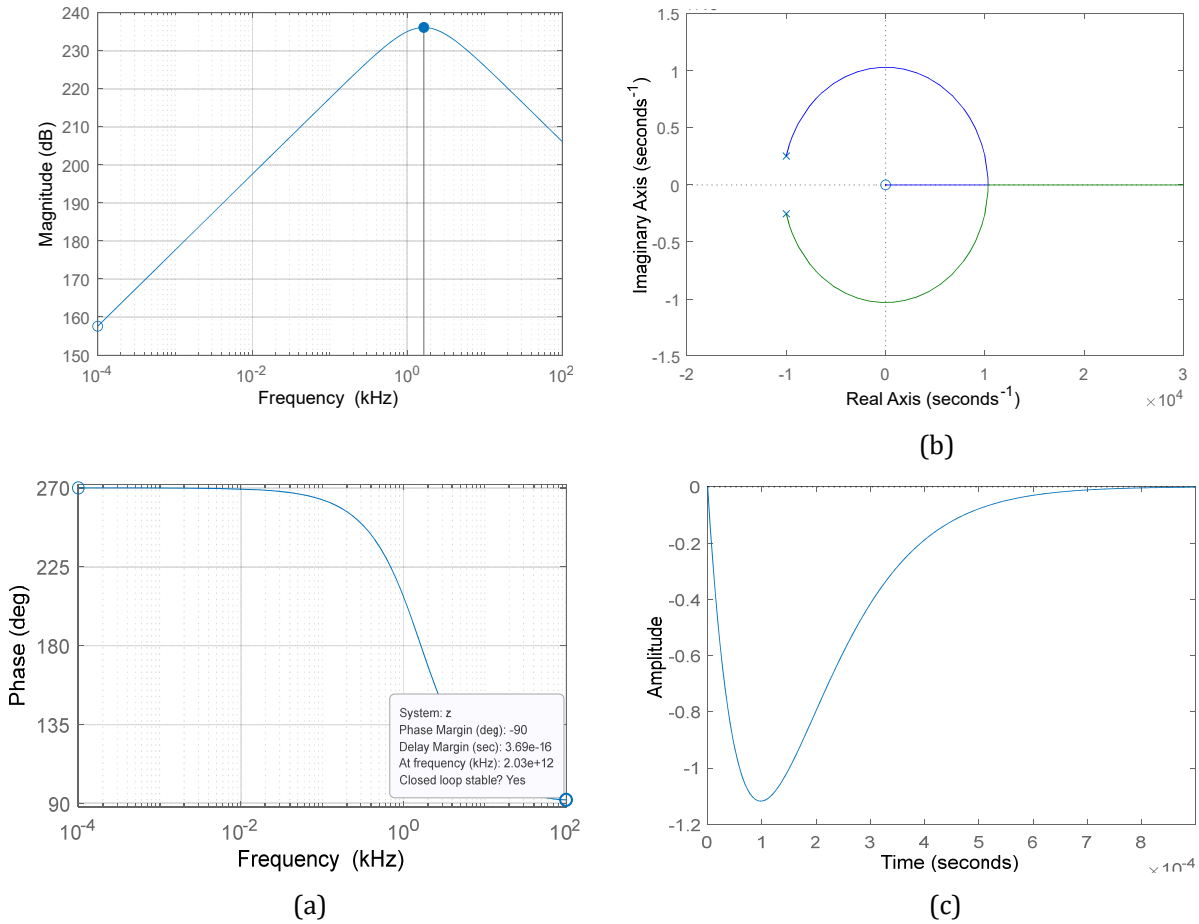


Figure 4.14 Voltage plant closed loop response with unity gain (a) Bode response (b) root locus (c) step response

The location of zero shows indirect proportionality to the system's inductance, that is, the larger the inductance, the lower the location of the zero; one of the major reasons lower inductances are preferred in boost converters. The frequency location of the zero shows direct proportion to the converter's load i.e., under lighter loads the location of the zero is closer to the origin on a root locus, whereas under heavier loads the location is further from the origin. From this, the control system must be designed such that the voltage loop bandwidth is lower than the RHP zero frequency location under the lightest anticipated system load conditions, for optimal stability. Load conditions are observed under the voltage

loop, as illustrated by in Figure 4.14 (c). The load in this study is a Li-ion battery; [17] showed that the variation of the Li-ion battery's impedance decreases with increasing frequency. The Li-ion battery's impedance is evaluated at $1kHz$ because of its characteristics; thus, at $1kHz$, each of its 4 cells in series have an impedance of $3m\Omega$, as per the research [17]. From this, the voltage loop bandwidth was determined using, f_z . It is worth noting that the loop bandwidth obtained was lower than, $1kHz$; hence, the largest excitation frequency for the Li-ion battery load was equivalent to the loop bandwidth frequency.

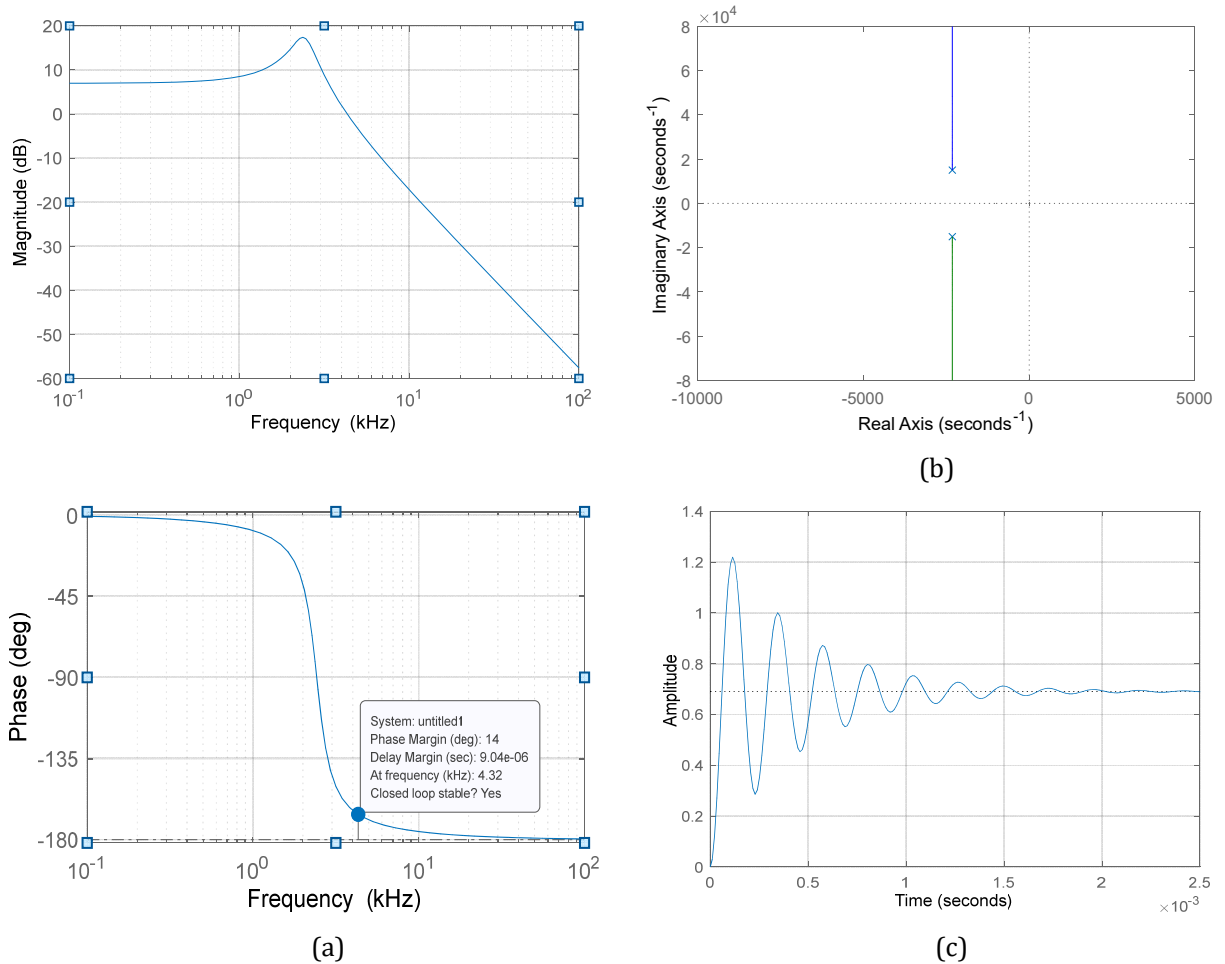


Figure 4.15 Current plant closed loop response with unity gain (a) Bode response (b) root locus (c) step response

Thus, an iterative method was used to determine the converging value for the loop bandwidth, as specified in Table 4.3. The bode plot shows the second order frequency response of the plant. The plot also shows that the gain margin is $-236dB$ – this is maximum gain the system can be reduced by before the system becomes marginally stable, and the phase margin is, -90° from RHP zero as discussed, which causes the observed delay in the output response. The bandwidth achieved by the loop is at $140Hz$ which does meet the desired specifications, however at a large crossover frequency. The peak response also needs to be reduced as it may amplify frequencies above the linear range during impedance measurements. The time step response shows that the system is underdamped. The settling time observed is $2.5ms$ which is within the specification required. Because the plant is non-minimum phase, it

experiences an undershoot of 71% - a value above the desired specification. With the zero close to the origin, the system gain is close to zero, thus the step response is always zero as observed.

The 3 plots for the current loop show that the closed loop is stable. The system has two open loop poles that are oscillatory. With a closed loop gain of 1, the closed loop poles sit out the desired region, hence a controller must be designed to ensure the system meets its specifications. The bode plot shows the second order frequency response of the plant. The plot also shows that the gain margin is infinite, and the phase margin is 14° - which lies outside the specifications. The bandwidth achieved by the loop is at 3480Hz which does not meet the desired specifications. The peak response also needs to be reduced as it may amplify frequencies above the linear range during impedance measurements. The time step response shows that the system is underdamped. The settling time observed is 2.5ms which is within the specification required. The plant experiences an overshoot of 71% - a value above the desired specification. The steady state error observed is 30%, which is not within the desired specification.

The loop plant is a *TYPE 0* system, meaning that it will have finite steady state error if given a step input. Using the final value theorem, this finite steady state error, $e_\infty(s)$, can be evaluated.

$$e_\infty(s) = \lim_{s \rightarrow 0} s \frac{1}{1 + P_{pv}^i(s)} r(s) \quad (4.74)$$

Since the input is a step function, its Laplace transform is:

$$r(s) = \frac{1}{s}$$

Substituting this transform, as well as (4.72) into equation (4.74):

$$e_\infty(s) = \lim_{s \rightarrow 0} \frac{s^2 + \left(\frac{\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}}{L} + \frac{1}{\tilde{\Delta}z_{bat}C} \right) s + \left(\frac{\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}}{\tilde{\Delta}z_{bat}LC} + \frac{(1 - D_{dc})^2}{LC} \right)}{s^2 + \left(\frac{\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}}{L} + \frac{1}{\tilde{\Delta}z_{bat}C} \right) s + \left(\frac{\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}}{\tilde{\Delta}z_{bat}LC} + \frac{(1 - D_{dc})^2}{LC} \right) + \frac{V_c}{L} s + \frac{V_c}{\tilde{\Delta}z_{bat}CL} + \frac{I_L(1 - D_{dc})}{LC}} \quad (4.75)$$

Evaluating the limit gives:

$$e_\infty(s) = \left[1 + \frac{\frac{V_c}{\tilde{\Delta}z_{bat}} + I_L(1 - D_{dc})}{\left(\frac{\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}}{\tilde{\Delta}z_{bat}} \right) + (1 - D_{dc})^2} \right]^{-1} \quad (4.76)$$

Which is a finite steady state error less than unity. Substituting for values evaluates to a steady state error of 0.35 which is 35% of the final value. This is close to the error evaluated in Sisotool, thereby affirming the loop error.

4.9.3 Controller design

Having analysed the closed loop of both plants, the controllers were designed to ensure that the closed loop system meets its specifications. MATLAB Sisotool was used to graphically design a controller that ensured that the system met the desired specifications.

i. Inner current loop

The inner current loop compensator was first designed. Because the steady state error isn't required to be zero, an integrator was not needed. This also helps improve the dynamics of the response since the integrator takes away 90° of phase from the loop. As such, a lead compensator, G_c was designed to increase the bandwidth while improving the steady state error [18].

$$G_c = \frac{a_c \tau_c s + 1}{\tau_c s + 1} \quad (4.77)$$

The parameters a_c and τ_c were designed for considering the specifications. Based on the figure, the lead compensator needs to add phase between $1kHz$ and $100kHz$, which will add some system gain, while increasing the bandwidth. The amount of phase to be added, is 60° at $40kHz$. To determine the parameters of the lead compensator, the following equations were considered. The maximum phase [18]:

$$\varphi_{max} = \sin^{-1} \left(\frac{a_c - 1}{a_c + 1} \right) \quad (4.78)$$

The maximum phase substituted was 90° plus an extra 15° degrees added as part of a lead circuit design safety factor. After obtaining a_c , and using the fact that this phase will be added at $40kHz$,

$$\omega_m = \frac{1}{\tau_c \sqrt{a_c}} \quad (4.79)$$

The parameter τ_c was solved for. In equation (4.79) [18], ω_m is $40kHz$. Substituting results from equations (4.78) and (4.79) into (4.77), the lead compensator obtained was:

$$G_c = \frac{21 \times 10^{-3} s + 1}{67 \times 10^{-4} s + 1} \quad (4.80)$$

The compensator designed was in fact, a lead compensator since $a_c > 1$. After simulating the closed loop, the system showed a slight lag in speed and the proportional controller gain was designed to increase the speed. The root locus was used to design for this. The root locus characteristic equation can be described from [19]:

$$P = \frac{N(s)}{D(s)} \quad (4.81)$$

The characteristic equation is:

$$D(s) + G_c N(s) = 0 \quad (4.82)$$

Therefore, substituting in $P^*_{pv}(s)$ into evaluates to:

$$s^2 + \left(\frac{\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}}{L} + \frac{1}{\tilde{\Delta}z_{bat}C} \right) s + \left(\frac{\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}}{\tilde{\Delta}z_{bat}LC} + \frac{(1 - D_{dc})^2}{LC} \right) + G_c \left(\frac{V_c}{L} s + \frac{V_c}{\tilde{\Delta}z_{bat}CL} + \frac{I_L(1 - D_{dc})}{LC} \right) = 0$$

The closed loop pole positions to achieve the desired loop speed while meeting the overshoot specifications are:

$$s = -154.5 \pm 101.1i$$

These closed loop pole positions lie at the edge of the desired region. The proportional gain required to place the closed loop poles at the desired location is thus:

$$\overline{G_p} = 2.329$$

This gain was then added to the loop and following results were obtained. (a) now shows that the $-3dB$ bandwidth of the system has reached $49.5kHz$ – with the peak response now minimised, as desired. The response now shows a first order response; this is because the location of the lead circuit pole is on the real line and closer to the origin than the system's open loop oscillatory poles, thus showing dominance. The bandwidth observed is lower than the specified $50kHz$, however, a further increase would result in a significantly large overshoot at the output as the system response gains speed. This does however, mean that frequency components above, $50kHz$ are expected to be attenuated by a gain of as much as 0.56. It should be noted that this is the attenuation of the induced excitation signal at the output of the solar module due to the converter's frequency response. The system now experiences a settling time of $4ms$ and a steady state error of 6.5% - within the desired specifications.

By considering Figure 4.17, the assessment of the attenuation experienced by the response signal was undertaken. This assessment is important because during experimental verification, there is a limitation in the smallest signals that the data acquisition equipment can pick up. That is, if the response signal is significantly attenuated, it might be too small for the equipment used; thus, understanding this attenuation will help choose the most appropriate data acquisition system. During the first stage, the figure shows that when an impedance spectroscopy duty cycle modulating signal is injected into the converter, the frequency response of the converter affects the amplitude of the induced excitation signal at the output of the solar module, as mentioned earlier i.e., the converter's frequency response is the input output relationship between the duty cycle and the induced excitation signal. The second stage now shows the solar PV module's AC EEC response – which is the input output relationship between the excitation and response signals for the module. Because the module's AC EEC is a passive low pass LC filter, the response signal will not experience gains larger than unity and will thus either equal the excitation signal in magnitude, for lower frequencies, or be attenuated for higher frequencies, as illustrated by Figure 4.18.

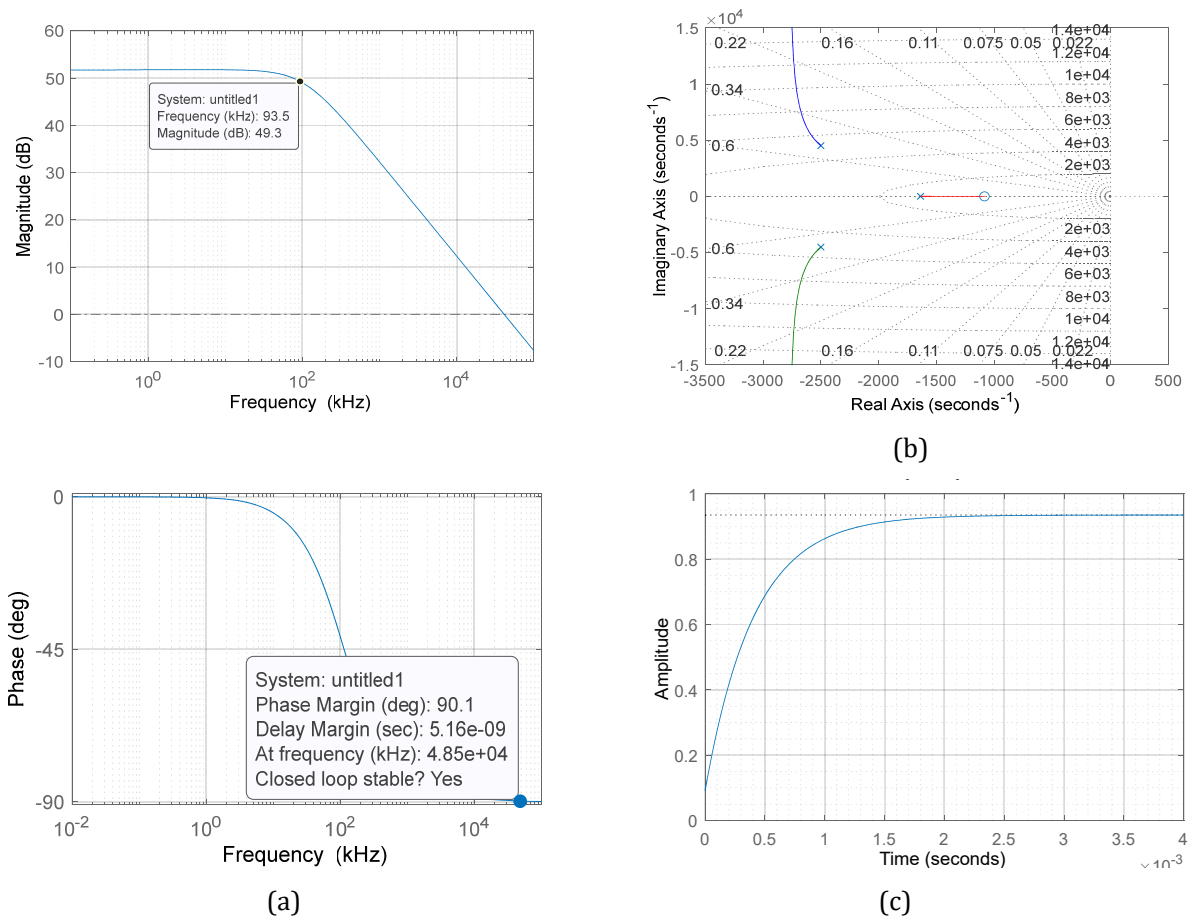


Figure 4.16 Lead compensated current loop response (a) Bode response (b) root locus (c) step response

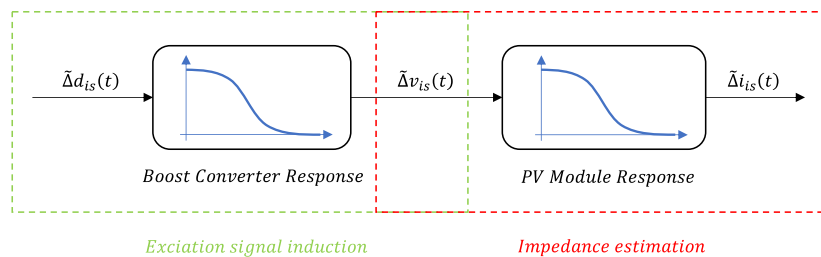


Figure 4.17 Stages in the induction of the excitation and response signals

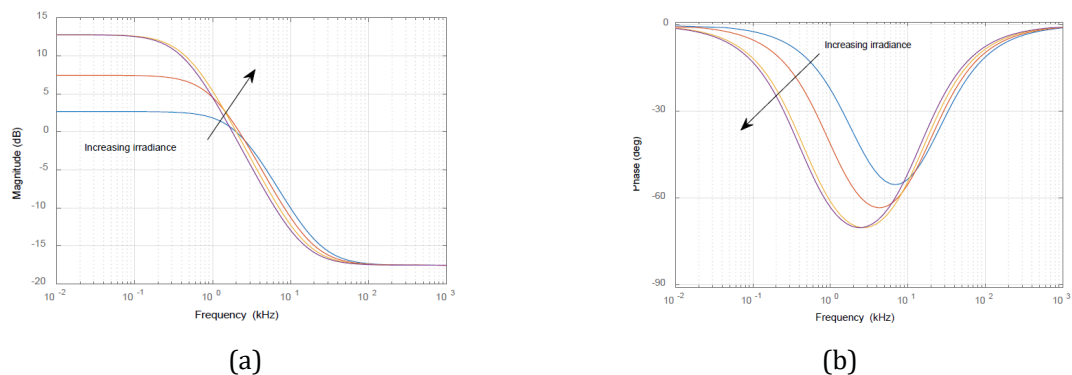


Figure 4.18 Solar module AC EEC frequency response under several operation conditions (a) gain (b) phase

The results were simulated using (2.8) and the parameter values extracted the solar module, as will be shown during the experimental verification. At the maximum impedance frequency of $100kHz$, response signal will experience an attenuation an attenuation of $-20dB$, an attenuation of 0.1, for all operational conditions; thus, this is the worst-case gain for the response signal. The determination of the smallest AC current value the data acquisition system will need to be capable of achieving will be based on this value. The figure also shows that the maximum phase difference between voltage excitation signal and current response signal can be expected around the between $100Hz$ and $1kHz$. However, at frequencies less than $10Hz$ and larger than $50kHz$, the excitation and response signals will be approximately identical in phase. This is important to note when performing the experimental verification because if the phase error introduced by the practical equipment is larger than the maximum achievable phase by the AC EEC, the accuracy of impedance measurements at lower and higher frequencies will be impacted.

ii. **Outer voltage loop**

The voltage loop requires a compensator to reduce its oscillations and overshoot – this a simple proportional plus integral compensator was designed.

$$G_v = \frac{sk_i + k_p}{s} \quad (4.83)$$

Where, k_i and k_p are the integral and proportional gain, respectively. To design the compensator, an integrator was added to the system, this makes the system Type 1, meaning it will achieve zero steady state error to a step input.

The next step was to place the zero at the desired location on a root locus such that the value of $(-k_p/k_i)$ could be determined. Based on the root locus, the desired location for the zero was, $s = -2300rad.s^{-1}$. The open loop gain was then adjusted such that the closed loop poles were within the optimal location within the design region as shown by Figure 4.19 (b). The response now shows an overshoot of 15% which is within the required specifications. the system has however, slowed now settling only in $2.5ms$; this is still within the required specifications.

The PM is now within the desired specifications at 92.5° . The system now shows a damping of 0.64, significantly reducing the oscillations. The bandwidth now sits at $189dB$ – still within the desired specifications – however, now at a much lower gain crossover frequency, as desired. Thus, the open loop now meets its required specifications.

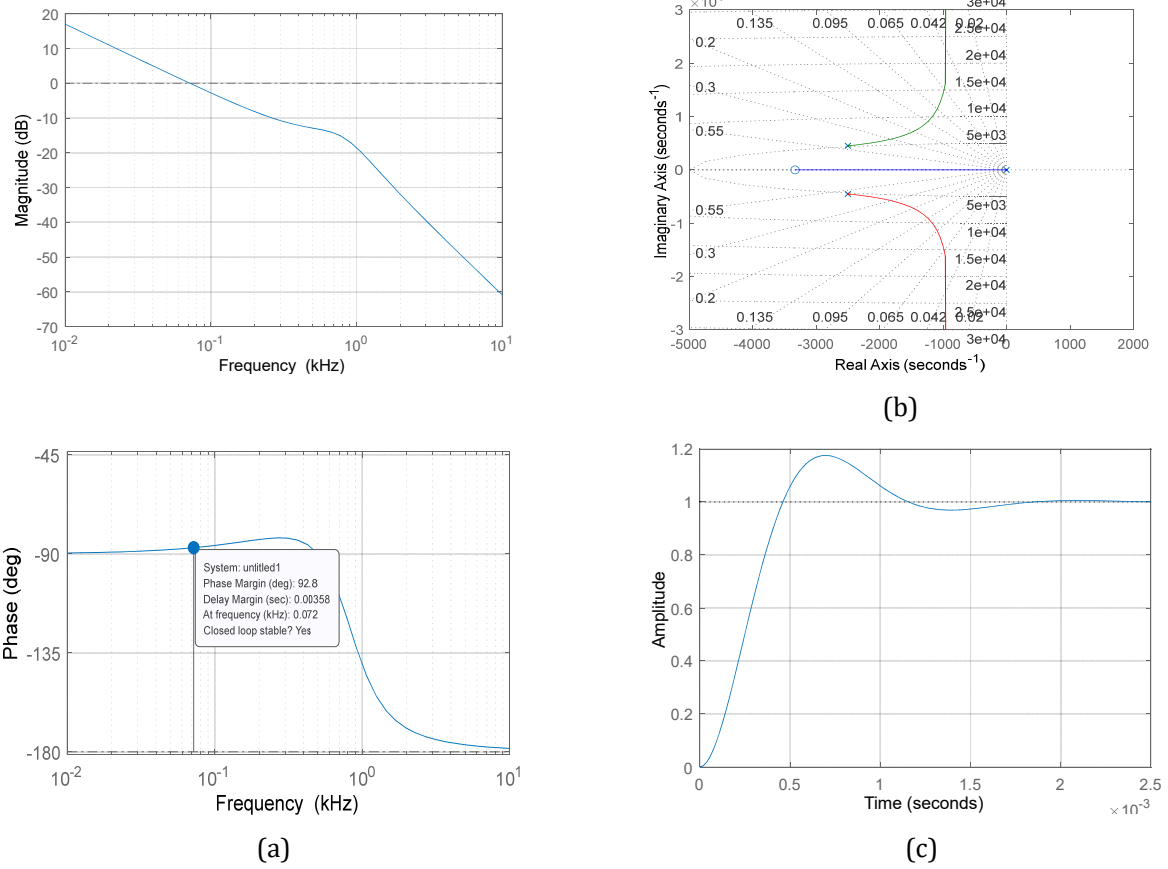


Figure 4.19 PI compensated voltage loop response (a) Bode response (b) root locus (c) step response

4.10 Maximum power point tracking control design – 2nd tier control design

The 2nd tier control subsystem required for optimal impedance measurements is the MPPT control; this enables the ECU to accurately determine the region in which the signal would be injected for device excitation. The control needs to achieve 2 major objectives: fast convergence in rapidly varying conditions and minimal oscillations at MPP, for a wider range operating points for signal injection. The perturb and observe (P&O) tracking algorithm is the most widely used for tracking the maximum power point for solar modules, this is because of its simplicity and ease of implementation. The disadvantages of the P&O algorithm include oscillations of the operating point around the maximum power point and the fact that as the irradiance decreases, the P/V curve flattens out [20]. Although the oscillations around the maximum power point can be reduced by decreasing the perturbation size, this, however, significantly slows down the algorithm's convergent time.

This work thus, implemented the incremental conductance algorithm to keep track of the maximum power point. This algorithm tracks the maximum power by comparing the instantaneous conductance of the module with its incremental conductance. The algorithm is superior to various others because it characterizes the module such that the exact perturbation direction can be determined [21]. More

importantly, it has a good convergence time under rapidly varying conditions – a vital characteristic in ensuring accurate perturbation amplitudes.

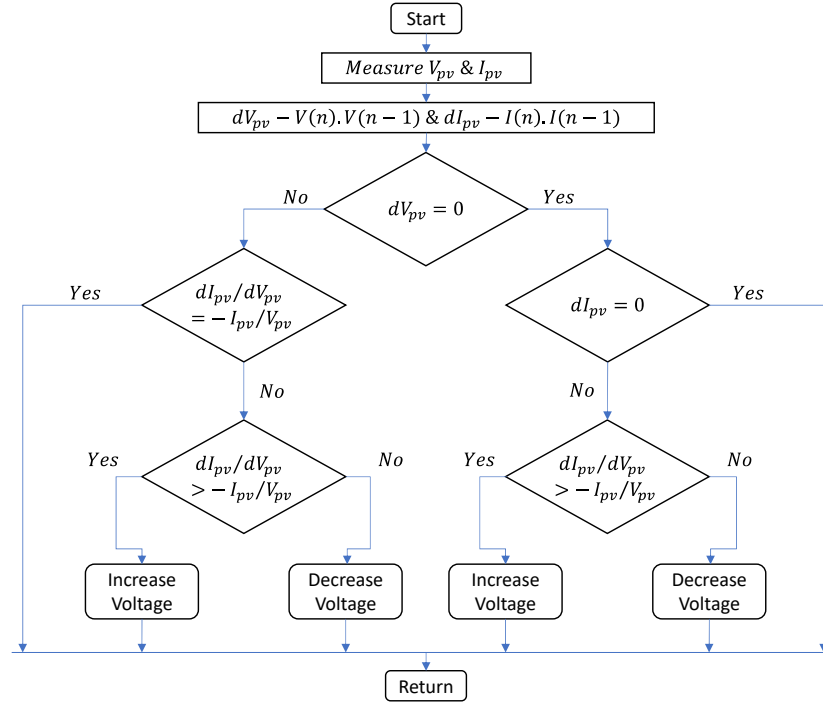


Figure 4.20 Incremental conductance maximum power point tracking algorithm flowchart for the module

To satisfy the condition of maximum power, the incremental conductance, $\partial P_{pv}/\partial V_{pv} = 0$, that is, at maximum power, the rate at which the module's power changes with respect to its voltage is null. The module's power given by: $P_{pv} = I_{pv}V_{pv}$; implicitly differentiating this power with respect to its voltage yields: $\partial P_{pv}/\partial V_{pv} = I_{pv} + V_{pv} \partial I_{pv}/\partial V_{pv}$ – the basis for the algorithm, as illustrated by Figure 4.20. The algorithm tracks the maximum power point noting that from the left of both I/V and P/V curves, the module's instantaneous conductance is greater than the module's incremental conductance, and the rate at which the modules power changes with respect to its voltage is negative i.e., $\partial I_{pv}/\partial V_{pv} < -I/V$ and $\partial P_{pv}/\partial V_{pv} < 0$, respectively. On the other hand, while the operating point is on the right of the maximum point, the module's instantaneous conductance is less than the module's incremental conductance, $\partial I_{pv}/\partial V_{pv} > -I/V$ and $\partial P_{pv}/\partial V_{pv} > 0$.

4.11 Impedance spectroscopy duty cycle perturbation signals' design

The perturbation duty cycle was designed to be small enough such that when added to the DC point duty cycle, the induced voltage excitation signal does not cause system to deviate from its DC operating point; this will ensure linearity in the measurements, as discussed in Chapter 3. As explained, the voltage excitation signal is first induced through an injected duty cycle, then the solar module responds to the induced voltage excitation signal – thus, leading to impedance estimation. This section thus, details the design on these duty cycle perturbation signals.

4.11.1 Excitation signal design specifications

The specifications for the excitation signal are detailed in Table 4.4. The perturbation signal is only injected once the system has reached steady state – in this case, steady state is defined as output voltage values that are within $\pm 10\%$ of $\overline{V_{pv}}$ – as discussed during the steady state evaluation of the control system design.

Table 4.4 Induced voltage excitation signal design specifications

<i>Specification type</i>	<i>Specification</i>
<i>Voltage Excitation Amplitude</i>	8.51% – 13.45% of $\overline{V_{pv}}$
<i>Measurement Frequency</i>	10Hz – 100kHz
<i>Testing Temperature</i>	25°C – 45°C
<i>Crest Factor</i>	< 1.5
<i>Data acquisition time</i>	< 10s
<i>SNR</i>	> 40dB

The voltage excitation amplitude is the range of amplitudes from the AC signal that will keep the system within its linear range of operation. The measurement frequency was determined by observing the Nyquist frequency response of the module when using an FRA to perturb the device – this will, again, be discussed in more detail during the experimental verification. The testing temperature was based on the safest achievable temperature under indoor laboratory conditions with the available equipment. Each of the excitation signals mentioned were thus designed based on these specifications, later simulated in this chapter, and experimentally verified in Chapter 5, to assess the feasibility of the signals for linear device excitation. All, except triangular excitation, are a form of induced excitation i.e., the voltage excitation signal is induced from an externally generated perturbation duty cycle, as will be analysed.

4.11.2 Chirp perturbation signal design

The design of a chirp perturbation signal is, perhaps, the simplest. In the time domain, the signal has a constant amplitude throughout its run-time, thus not requiring improvements in its crest factor, which is generally, 1.45 [24] – within the desired specification. When designing the signal only 4 parameters are required, namely: the amplitude, start and stop frequencies (which determine whether the signal will be an up-chirp or down-chirp signal), and the frequency distribution. The amplitude is designed by considering the “*voltage excitation amplitude*” specification – particularly the signal amplitude that would lead to the maximum “*voltage excitation amplitude*” – i.e., 13.45% of $\overline{V_{pv}}$. From (3.9), (4.23), assuming negligible parasitic resistances and input current ripple, the injected chirp perturbation duty cycle is:

$$\tilde{d}_{in} = \left[\left[1 - \left(\frac{\overline{V_{pv}} + \tilde{\widehat{v}}_{in}^m}{\overline{V_{bat}}} \right) \right] - D_{dc} \right] \cos(2\pi f_{is}(t)t + \omega_o) \quad (4.84)$$

Where, $\tilde{\Delta}\hat{v}_{in}^m$ is the maximum voltage excitation amplitude. The start and stop frequencies are based on the "*Measurement Frequency*" range specification from Table 4.4. Lastly, the frequency distribution can either be linear or geometric; a choice which is dictated by the module's operational conditions – more on this during experimental verification. During simulation results, the design will be assessed accordingly.

4.11.3 *Triangular wave excitation signal design*

This form of excitation is the least researched and implemented. This is because it is not as easy to generate as a square wave, as well as the fact that it does not provide fundamental frequency characteristics that are as good as a pure sine wave. When using a boost type SMC as an excitation source, however, this type of excitation presents a unique opportunity for device excitation. The inductance in series with the DUT in the SMC forces the input current to increase and decrease gradually, as discussed during the SMC's operational analysis; this creates an input current and hence, input voltage, that is a triangular waveform. These are small high frequency switching signals that are generally undesired during converter-based impedance spectroscopy measurements.

In this type of excitation, however, the signals can be used in direct excitation, contrary to induced excitation resulting from an externally injected perturbation duty cycle. The amplitude design of the signal is based on (4.3), while the signal definition was discussed in (3.7) and (3.8). To control the excitation amplitude, while keeping the system within its linear range, the frequency of the converter was varied according through a control technique referred to as foldback control.

4.11.4 *PRBS perturbation signal design*

The PRBS perturbation duty cycle signal undergoes multiple optimizations during its first design level, as will be discussed, to ensure optimal bandwidth characteristics in the induced voltage excitation signal.

i. *Frequency span*

Based on the Wiener-Khintchine theorem – which states that the density of the spectrum resulting from a stationary random process is the Fourier transform of its auto-correlation function – the spectrum of the signal can be shown be bandwidth-limited such that the usable frequency band is approximately as shown by (4.85) [24].

$$f_{band} \leq 3f_c \left(\frac{1}{3} - \frac{1}{N} \right) \quad (4.85)$$

Where, f_c , is the corner frequency of the spectrum and N is the number of terms in the bit sequence. The minimum and maximum frequencies that was obtained from the spectrum are approximately:

$$\left. \begin{aligned} f_{min} &\geq \frac{1}{N\Delta t} \\ f_{max} &\leq \frac{1}{3\Delta t} \end{aligned} \right\} \quad (4.86)$$

Where, Δt is the clock pulse duration.

ii. Signal to noise ratio

The smaller the bit width, the closer the PRBS signal's autocorrelation function is to a delta function, which would result in more accurate PV module impedance measurements. But a smaller width also results in a lower SNR. A small SNR is non ideal for impedance extraction as it would introduce significant phase error in the impedance. The bit width was therefore chosen empirically such that the signal could be accurate but with the highest SNR obtainable.

iii. M-sequence length

The choice in the M sequence length of the signal was designed such that:

$$M \geq \frac{t_s}{\Delta t} \quad (4.87)$$

Where, t_s , is the response time of the system. From, (4.87), in a PV impedance extraction system, the greatest time constant Δt_{max} , was used to determine the sequence length such that:

$$M = 5 \left(\frac{3}{2} \right) \Delta t_{max} \quad (4.88)$$

Where, the greatest time constant can be expressed as in (4.89) [24]:

$$\Delta t_{max} = t_{delay} (2^n - 1) \quad (4.89)$$

$$\varphi^p \left(m \frac{2\pi}{N t_{delay}} \right) = \frac{1}{N} \varphi^d(m) \left| \frac{\sin \frac{\pi m}{N}}{\frac{m\pi}{N}} \right|^2 \quad (4.90)$$

iv. Sequence repetition

The number were chosen to be as large practically allowable within a given period. This was to help eliminate the influence of white noise and other system noise. In doing so, the system exhibits a large SNR.

v. Amplitude

The last variable to optimize is the signal's amplitude. Fortunately, this follows both the triangular and square wave signals' amplitude design, such that:

$$\tilde{d}_{in}(t) = \left[1 - \left(\frac{\overline{V_{pv}} + \tilde{v}_{in}^m}{V_{bat}} \right) \right] - D_{dc} \quad (4.95)$$

This is the amplitude that ensures linearity in the induced voltage excitation signal, as discussed through several instances in the section. It is worth noting, however, that this equation specifically considers the amplitude of the perturbation signal, with the presumption that the converter's frequency

response has negligible effect on the induced excitation signal – the current research standard based on the literature reviewed.

4.12 Simulation results and discussions

Both the single phase and multiphase boost converters were simulated with MATLAB's Simulink environment. The simulations were executed with a fundamental fixed step size of 1×10^{-6} s, through the solver: *ode23t(mod.stiff/Trapezoidal)*. Both these models are illustrated in Appendix B. For the purposes of this study, the simulations were performed under closed loop compensation. The module was emulated as a voltage source with its equivalent impedance, while the load battery was emulated as a constant current source representing its variable charging current demand. A reference step with filtered high frequency components was used to reduce the demand on the control action on each of the controller designed. For the simulation results, the multiphase configuration was used as the default converter, unless stated otherwise.

4.12.1 Output voltage

The input current was simulated such that the module operates at its maximum power – under insolation conditions of, 800 W/m^2 and temperature of, 30°C – which is 17.55V and 3.61A ; this is a maximum power of 63.36W . The charging current was set to be 25% of the battery's capacity – a rate recommended by most manufacturers. Figure 4.21 (a) shows that the single-phase boost converter reaches an output voltage terminal value of 34.20V with a steady state error of -0.01V , in a settling time of 70ms ; this is a 0.03% error of the steady state value, within the design requirements. The transient response is faster than the reference signal which results in a 12% overshoot in the output – this still within the design requirements. Figure 4.21 (b) shows that the multiphase configuration achieves similar transients to the single phase, but with a higher overshoot of 19% – the higher overshoot can be attributed to the larger rise time from the configuration.

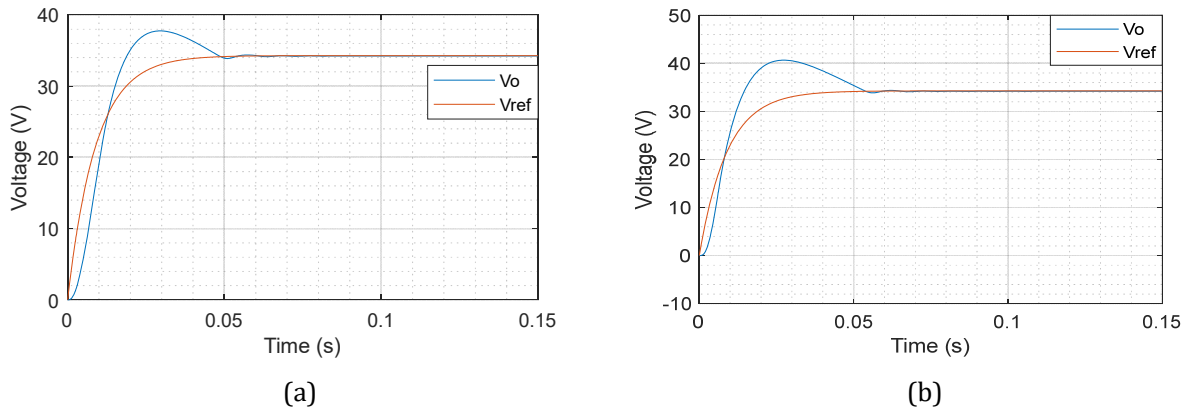


Figure 4.21 Steady state boost converter input and output voltage (a) single phase (b) multiphase

The smaller effective inductance value also contributes to the larger overshoot. The configuration reaches a terminal value of 34.11V – slightly lower than the single-phase boost because of the increased

voltage drop – with a steady state error of -0.09 , in a settling time of 80ms ; a 0.26% error in its steady state value, within the design requirements. It should be noted that, because of the increased voltage drop, the multiphase configuration reaches its required terminal value at a duty ratio higher than that of the single-phase counterpart. This will lower the efficiency of the multiphase configuration compared to the single-phase.

4.12.2 Output voltage tracking specification

The converter was assessed in terms of its ability to charge a 3 different kinds of battery series configurations, 12V , 24V , and 48V ; this was done to assess the ability of the control to track different voltage setpoints. To charge a 12V battery at the maximum recommended charging capacity of $0.25C$, the charging voltage needs to be approximately 18V – accounting for the voltage drops and battery's internal resistance; with charging voltages of 34V and 60V , for the 24V and 48V batteries, respectively. The charging voltage was monitored to observe the regulation at the output of the converter; the output voltage is illustrated by Figure 4.22.

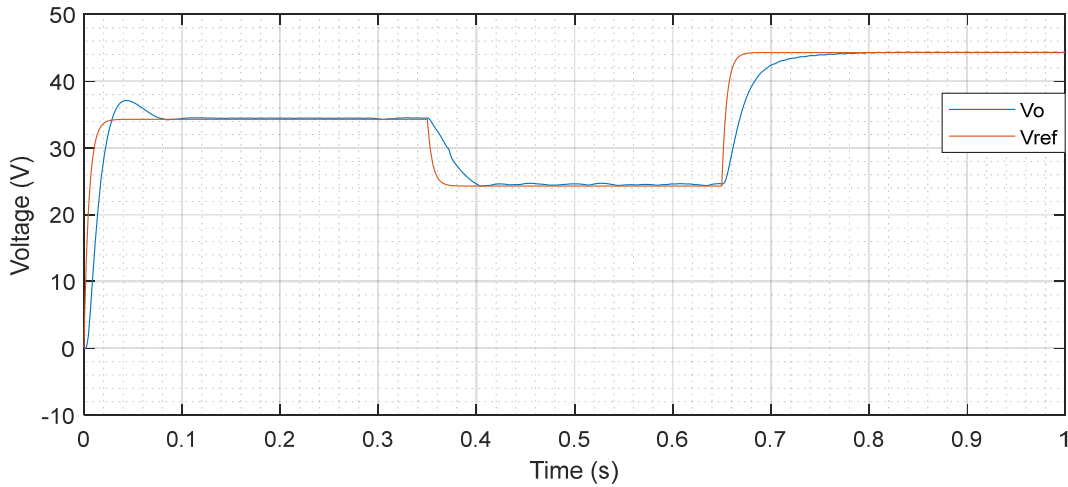


Figure 4.22 Charging voltage tracking specification

The results show that the control can track all the setpoints within the output voltage requirements. The 34V step has the same characteristics as discussed in the previous section. The 18V step shows the lowest achievable output voltage by the converter; this is dictated by the minimum duty ratio. The converter reaches a terminal voltage of 24.56V – a steady state error of $+2.33\%$. The larger error is largely due to the poor utilization at lower operating duty ratios. This error is, however, still within the design requirements and acceptable for charging the battery. It is important to note that this step is critically damped with a settling time of 65ms – thus meeting all the design requirements. The 60V step is shown to be critically damped, settling in 135ms – this is because of the larger duty ratio required to reach this voltage. Although it settles the longest, it is still within the requirements. The voltage reaches a terminal voltage of 59.11V – a steady state error of 0.89V , -1.48% .

4.12.3 Maximum power point tracking

With incremental conductance employed, as discussed earlier, the system ability to operate at maximum power was evaluated. The module was simulated to emulate maximum power points at both varying temperature and irradiances. Figure 4.23 and Figure 4.24 both illustrate these results. Figure 4.23 shows that as the irradiance increases, both the module's output voltage and current increase – with the current increasing more significantly than the voltage – which matches the theory development of solar modules. Maximum power for the simulated module is achieved when the power curve reaches steady state. It is evident that at lower irradiance values, the algorithm shows a marginally stable system with oscillations around the maximum power point. The convergence time at locating the maximum voltage is approximately 3s during lower irradiance values but improves significantly as the irradiance increases. The control does show significant current spikes at larger irradiance values, attributed to the operating duty ratio at these irradiance levels. Figure 4.24 shows better transient and steady responses by the algorithm. This is largely because the temperature change has little effect on the current, thus the demand on the control action is lower than during irradiance changes. The convergence time on the voltage is $< 1s$ which is good for rapidly varying conditions.

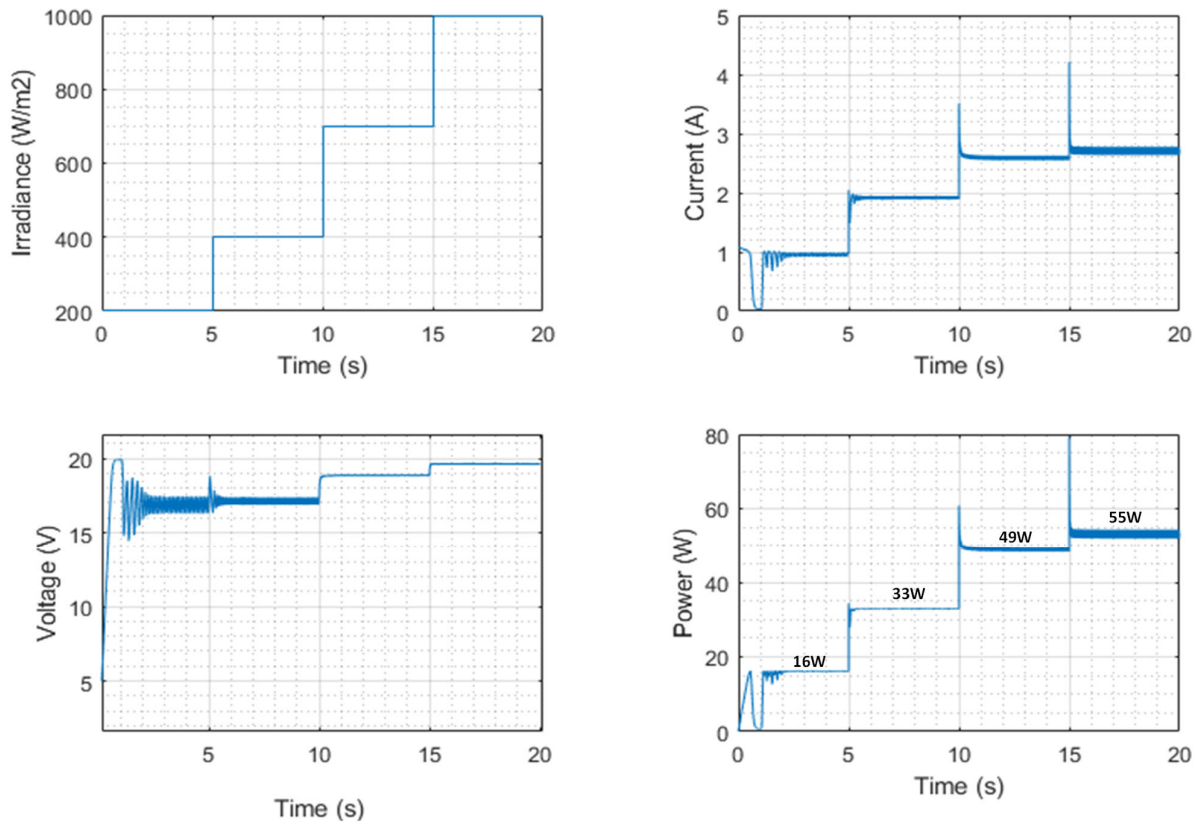


Figure 4.23 System's maximum power point tracking with varying irradiance

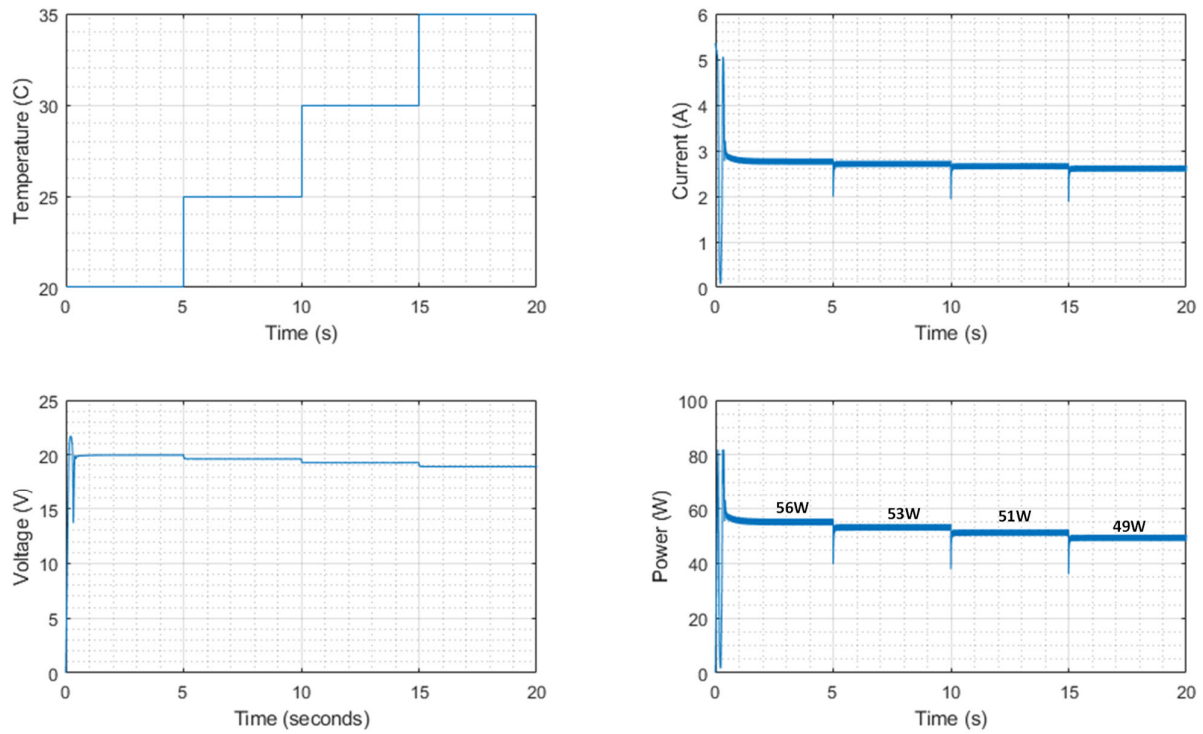
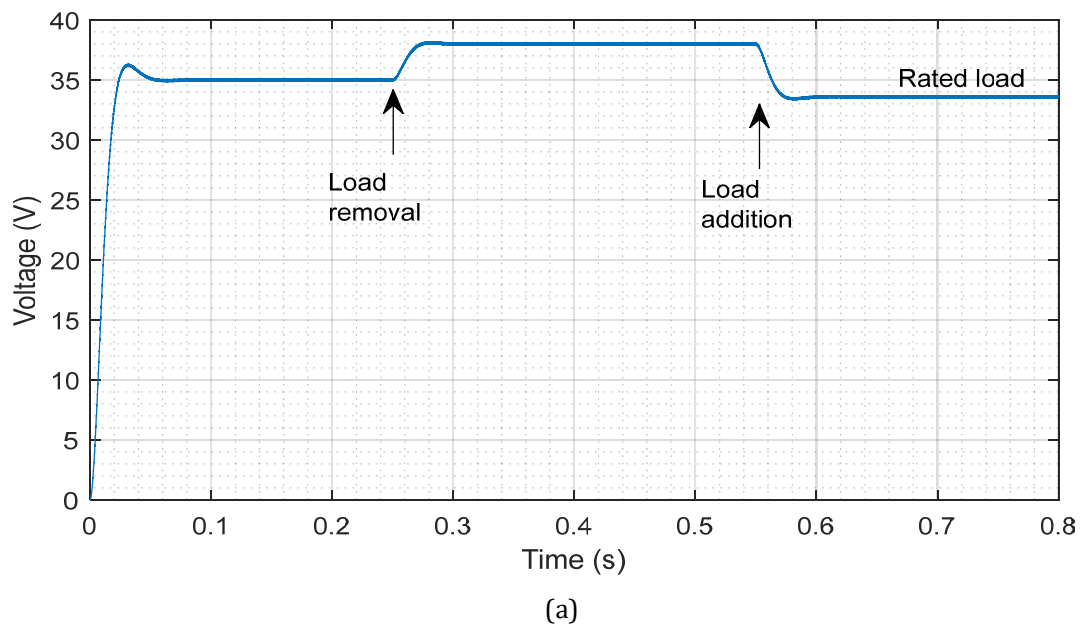


Figure 4.24 System's maximum power point tracking with varying temperature

4.12.4 Load step

The charging current of a battery is dependent upon the device's power needs. The system was tested on its ability to provide regulated charging voltage while the battery undergoes its pre-charge, constant current, and constant voltage states; This was simulated by using a controlled current source as a load. The commands set for the source were set to step up or down as illustrated by the figure. The charging voltage provided by the converter was observed.



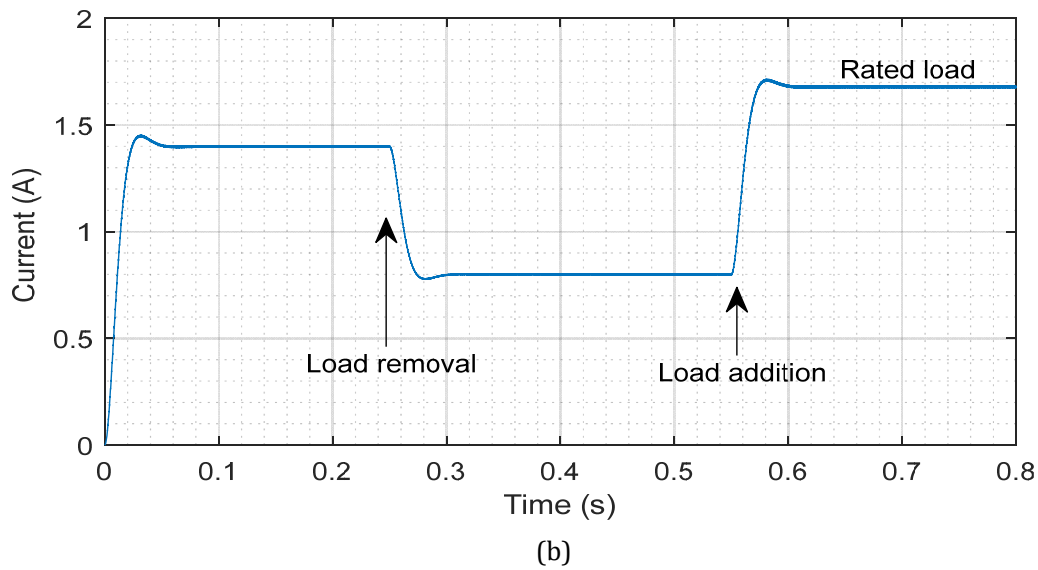
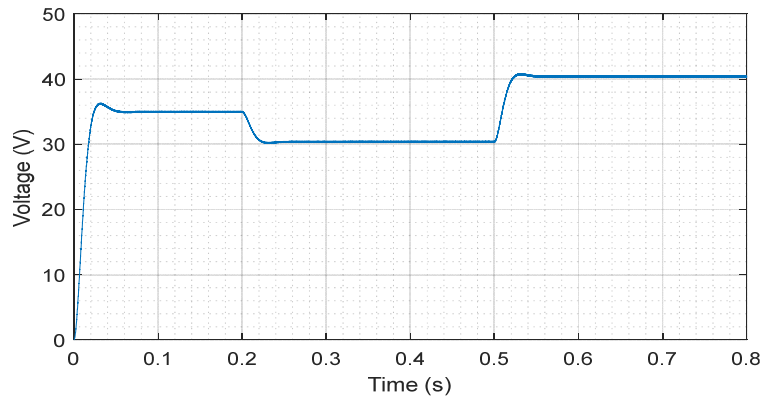


Figure 4.25 Load disturbance rejection (a) output voltage (b) output current

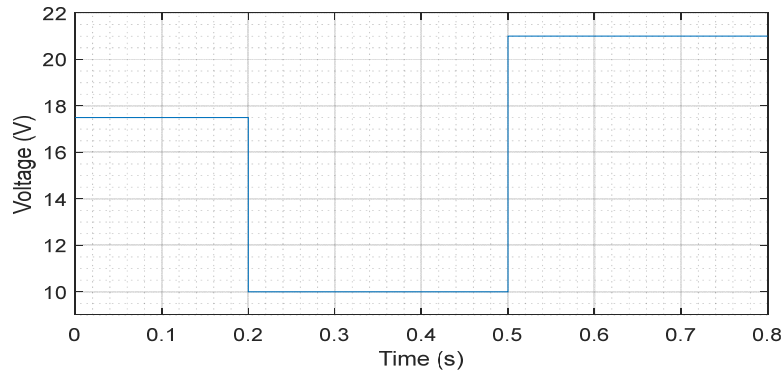
Figure 4.25 (a) shows that when the load changes as illustrated in Figure 4.25 (b), the charging voltage pivots away from its nominal charging voltage; this is largely because the control system was design to operate at a single operating point. It is worth noting during impedance measurements, the module's output voltage will be the concerning variable; that is, when the load changes, the input current demand changes, and based on where on the I/V curve the module is operated, the load changes may impact the module's output voltage. This will be further elaborated upon during experimental verification with a physical module being used. The steady state error during the load changes remains within $\pm 10\%$ as per the specifications. At rated load, the error is, -6% , which is sufficient in developing an adequate potential difference to charge the load battery.

4.12.5 Input voltage step

This test was performed to assess the ability of the converter to provide the required load voltage regulation while the module's output voltage changed. Based on the selected temperature values, the module's output voltage was stepped down from $17V$ to $10V$ at $200ms$; it was then stepped up from $10V$ back down to $21V$ at $500ms$. Figure 4.26 shows that the control system is able achieve good transient behaviour when the output voltage of the module varies from the converter's rated minimum input voltage and to its maximum rated input voltage. The control, however, fails to meet the steady state error specifications when the operating point system varies. At the minimum voltage, the steady state error observed is, 13% , while at the maximum input voltage, the error is, 16% . These errors show that performing impedance measurements at different operating points under the current control architecture may not be possible. This is dependent upon the induced excitation signal, as discussed earlier. During experimental verification, this will be assessed further.



(a)

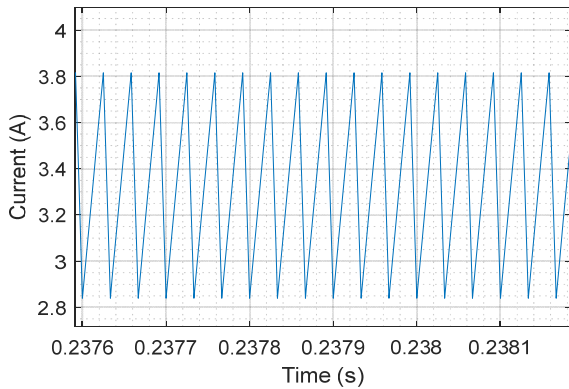


(b)

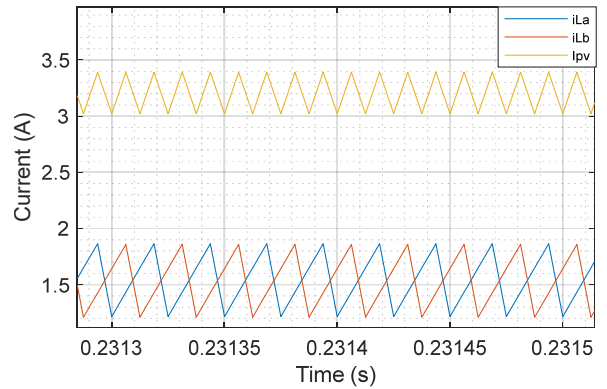
Figure 4.26 System input voltage disturbance test (a) output voltage (b) input voltage

4.12.6 Input current and output voltage ripple

The multiphase boost topology was specifically chosen to improve upon the converter's input current ripple for better signal processing at small signal excitation levels. Therefore, both single phase and multiphase converters were simulated to assess the performance of their current ripple at rated power and load.



(a)



(b)

Figure 4.27 Boost converter input current ripple (a) single phase (b) multiphase

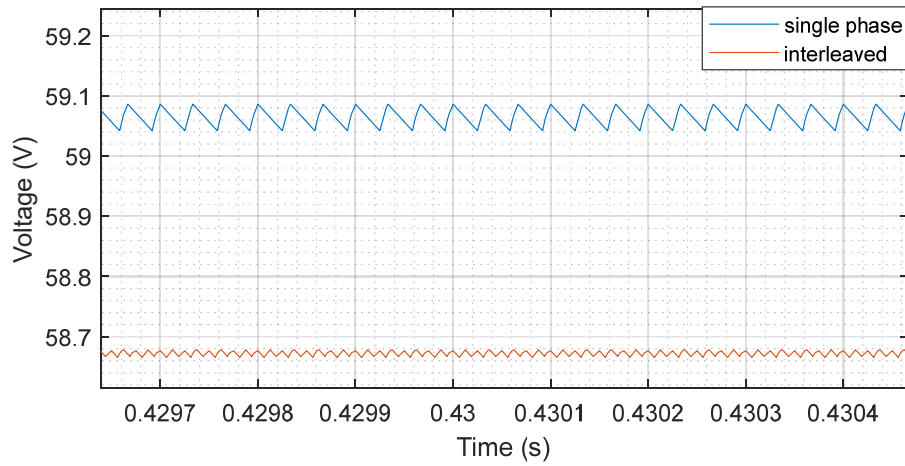


Figure 4.28 Boost converter output voltage ripple

Figure 4.27 (a) shows that the input current ripple is 0.9A peak-peak – which is 27% of the module’s output current – this is within the desired specifications. Figure 4.27 (b) shows that both inductor current ripples in inductors, L_a and L_b achieve a percentage of 39% with its average value. This is still within the recommended 40%. In input current ripple, however, is 12.5% of the module’s output current – as desired from the theory and requirements. This is less than half that from the single-phase converter. The load voltage ripple is 0.067% of its terminal value for the single-phase boost converter, while it is 0.033% for the multiphase, as illustrated by Figure 4.28.

4.12.7 Impedance spectroscopy voltage excitation signal induction

Now that the converter has been confirmed to operate as designed for, its use as an excitation source in the implementation of device impedance measurements in real-time was evaluated. The analysis was undertaken by assessing the induced voltage excitation signals from an AC perturbation modulating duty cycle signal, as illustrated by Figure 4.1. The simulation model for the solar module took the form of voltage source in series with the module’s first order AC EEC from a single operating point. The parameter values for the AC EEC were determined experimentally, as will be discussed in Chapter 5. Each perturbation was taken once the system had reached its steady state i.e., $\pm 10\%$ of the average charging load voltage. With each perturbation type, the properties of the induced voltage excitation signal were compared to the perturbation duty cycle signal, and subsequently analysed to assess the feasibility of device impedance estimation with the perturbation signal type. The main properties considered were, crest factor, induced voltage excitation gain, excitation bandwidth, SNR, and impedance acquisition time. It is worth noting that, simulation results have little to no noise, therefore, the high frequency inductor current ripple (considered EMI noise) will be assessed to determine the signals’ SNR. While the data acquisition time is dependent on the available digital signal processing (DSP) equipment, each of the signals will be compared while using the same DSP unit, hence, this will give a valid comparative analysis between the signals. It is worth noting that, although the power regulation system is operating under closed loop control, the injection of the excitation signals was implemented in open loop; this is

valid for implementations without a varying operating point. In Chapter 6, the closed-loop implementation of impedance estimation relating to varying operating points will be discussed in detail.

i. **Chirp excitation**

The chirp signal has various properties that make it an ideal excitation signal of choice. The most important property in the context of implementing broadband impedance spectroscopy being that it has a low crest factor. That is, its response in the time domain has a uniform magnitude across its time span. However, this only applies with direct excitation. When a DC-DC converter is utilized as an excitation source, a modulating duty cycle of the form in (3.9) results in a voltage excitation signal as illustrated by Figure 4.29. As seen from the figure, the 2 main issues with a non-parametric modulating signal, firstly, the injection of this type of signal results in large voltage overshoots during signal induction at low frequencies; secondly, it results in the induction of voltage excitation signals that contain large amplitudes at lower frequencies and small amplitudes at lower frequencies, thus reducing the optimal frequency range of the chirp excitation signal, thereby limiting the perturbation frequency range for the module.

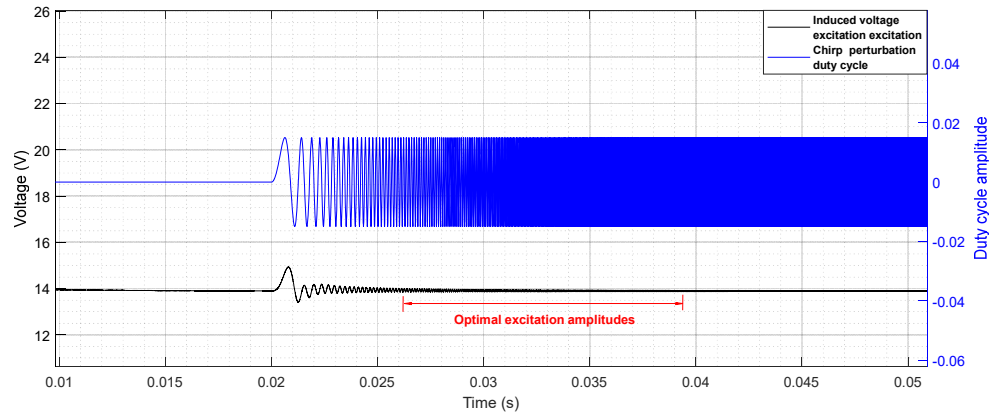
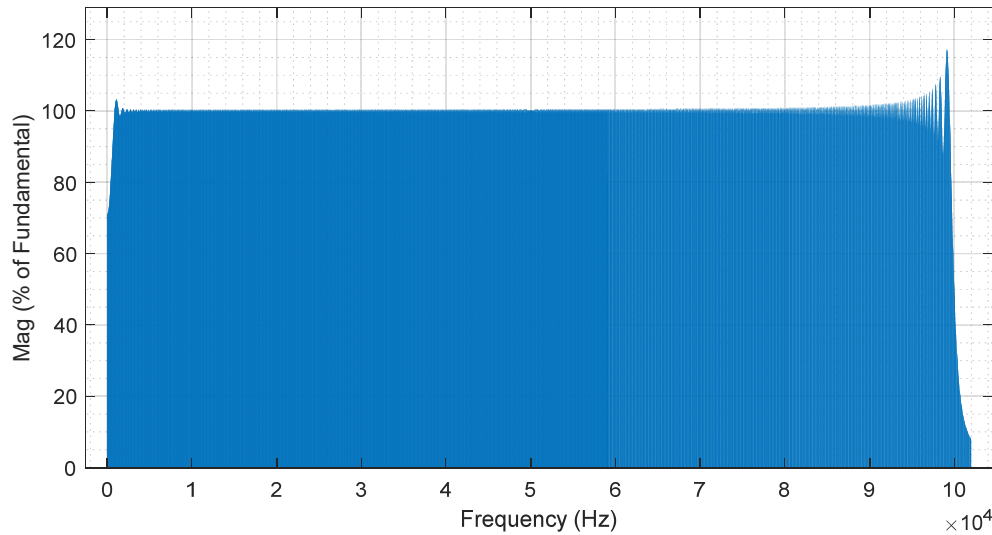


Figure 4.29 Chirp excitation analysis



(a)

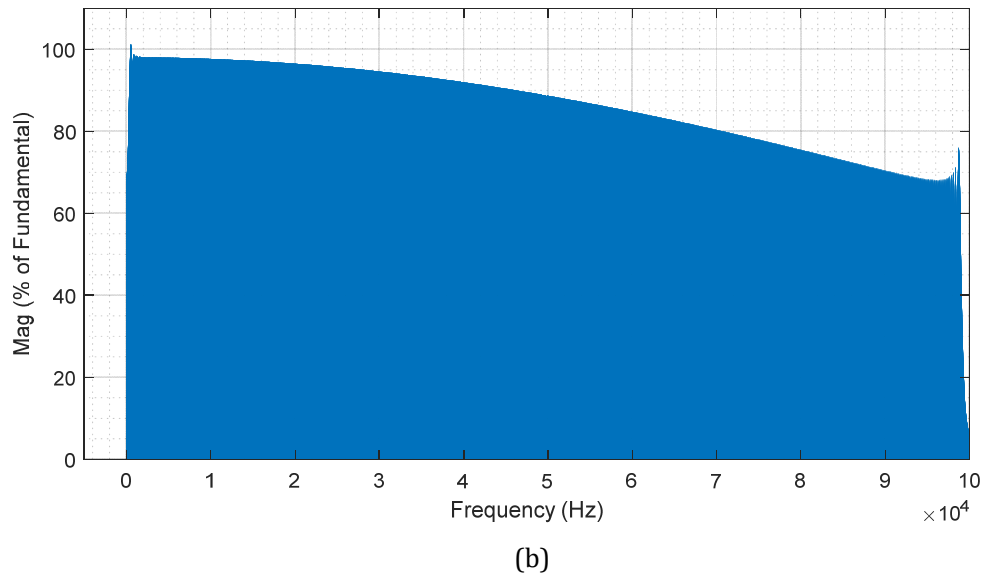
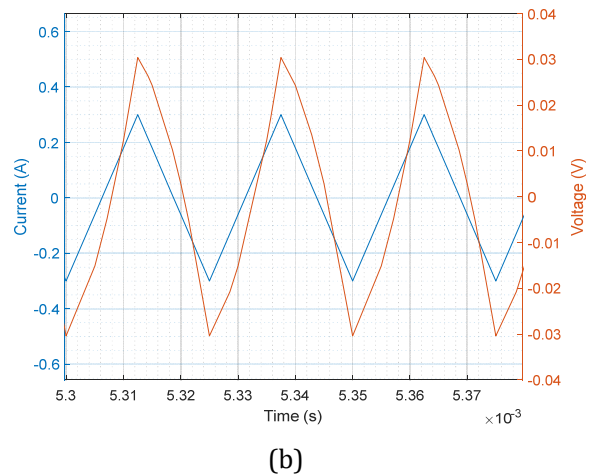
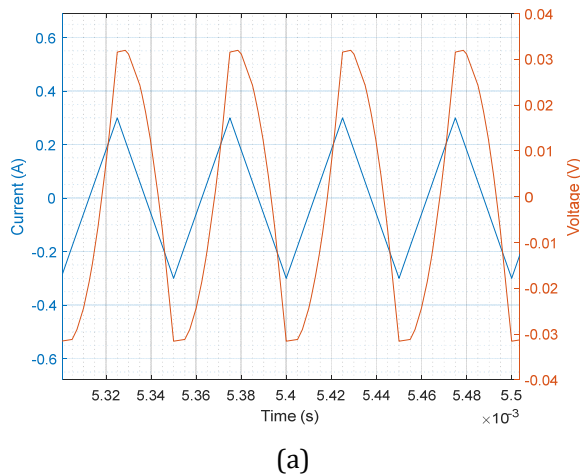


Figure 4.30 Chirp excitation frequency comparative analysis (a) perturbation duty cycle signal (b) voltage excitation signal

The modulating perturbation duty cycle signal thus, needs to be tailored to the frequency response of all operational points within the system's operational conditions. The crest factor of the voltage excitation signal is, 3.12 because of the large gain at low excitation frequencies. The SNR of the signal was, 90dB, while the data acquisition time was 4.32s – within the required specification. The frequency response of the duty and perturbation signals was analyzed and shown in Figure 4.29. Evidently from Figure 4.30 (a) the frequency response of the perturbation signal is a flat spectrum that contains the full spectrum of frequencies as designed for; Figure 4.30 (b), however, shows that the excitation signal's effective excitation bandwidth is, < 21.5kHz as a result of high frequency attenuation.

ii. *Triangular wave excitation*

As mentioned earlier, the converter's inductor current ripple was used as the triangular excitation signal, a form of direct excitation since no external perturbation duty cycle signal is used; as such, the response current signals were observed as well.



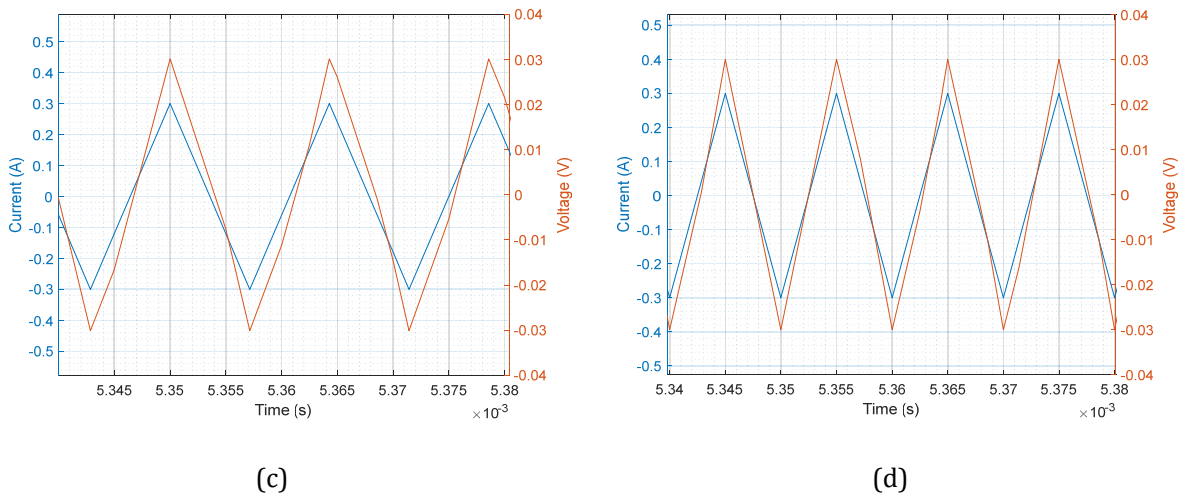


Figure 4.31 Triangular wave inductor current excitation and voltage response signals at varying frequency (a) 20kHz (b) 40kHz (c) 70kHz (d) 100kHz

The amplitude of the current ripple was designed such that it lies within the amplitude specified in Table 4.4. The frequency array of the signals was chosen as, $[20, 40, 70, 100]kHz$ – at a constant 50% duty ratio; this set would yield a good approximation under high irradiance conditions for the module, conditions which are more response to higher frequencies (more on this choice of frequencies during experimental verification and Chapter 9). Figure 4.31 illustrates the expected response signals by the module's emulated AC EEC. The symmetry does not appear to influence the response signals; however, the response signals show, as evident from Figure 4.18, that the AC EEC's frequency response under these conditions attenuates large frequencies at the same gain of $-20dB$ – there is no difference in magnitude from the voltage response signals at the tested frequencies. The lowest switching frequencies show the largest phase shift in the response signal, as observed from the change in signal shape – whereas higher frequencies exhibit little to no phase shift, as hypothesized through Figure 4.18, for the chosen operational point. It is worth noting that the decoupling capacitor will influence both inductor current and response signals, as will be discussed during experimental verification. The crest factor of the induced excitation signals ranges from 1.78 – 1.87, lying outside the specifications. The lowest SNR observed was 81dB, while the acquisition time was 5.57mins, not within the designed specifications. This is largely due to the fact that this is a single frequency excitation signal.

iii. **PRBS excitation**

The signal was generated on MATLAB Simulink. A 12-bit generator was used. Although a higher bit generator could be used to achieve high frequency resolution for better impedance estimation, this would increase the data acquisition time for the sequence. A lower bit generator could be used to achieve faster data acquisition times; however, this would result in lower frequency resolutions. Therefore, 12 bits is a trade-off that strikes a good balance between resolution and acquisition time. The minimum and maximum clock frequencies for the signal were chosen such that they were 5 times the minimum and maximum perturbation frequencies, respectively. For the PRBS signal, the quality factor of the converter was

designed to accommodate the smallest signal period. Figure 4.32 (a) shows the PRBS modulating duty cycle signal generated on MATLAB Simulink with the use of the generator. The signal shows a constant amplitude, 1.35 crest factor signal which is adequate for linear device perturbation. The signal has clear distinct low and high frequency bands at regular deterministic intervals.

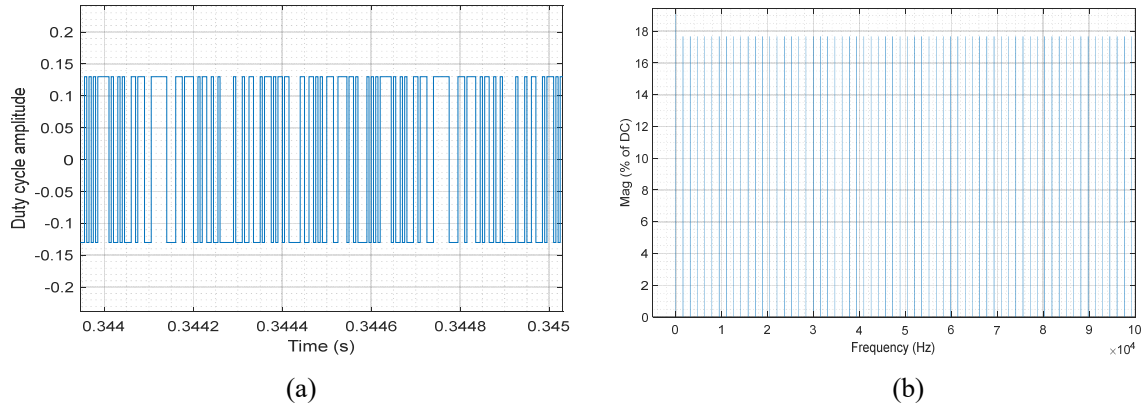


Figure 4.32 PRBS modulating duty cycle perturbation signal (a) PRBS time domain signal (b) perturbation signal frequency power spectrum

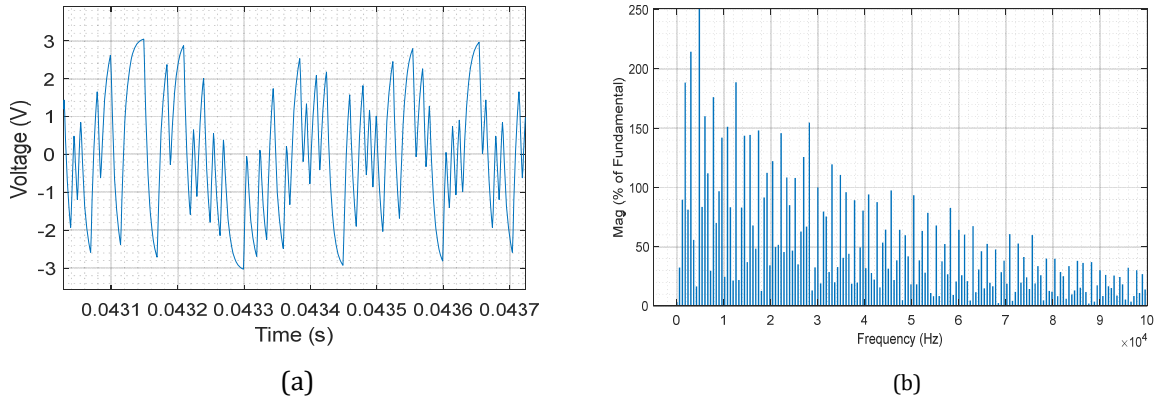


Figure 4.33 Induced voltage signal results (a) excitation time signal (b) excitation signal frequency power spectrum

Figure 4.33 (b) shows that the modulating signal's frequency response has a, $-3dB$ bandwidth of approximately, $99kHz$, which is sufficient for crystalline module excitation. Figure 4.33 (a) shows that induced voltage excitation signal now has slow rise and fall times, for reasons already discussed in the previous section. The signal shows that the at high frequency bands, the excitation signal reconstruction is poor because of the limit to the converter's quality factor i.e., it can only be made as large as possible, only within the design limits. Hence, Figure 4.33 (b) shows that the excitation signal has an excitation bandwidth of only, $47kHz$, less than the perturbation signal. The crest factor has increased to 3.1 – above the specifications. The acquisition time for the signal was, 3.78s, with a SNR of, $54dB$, within the desired specifications. The SNR is lower in the PRBS signal because the signal has white noise properties, this is well illustrated by Figure 4.33 (b). Because the signal provided the quickest data acquisition time, it is

desired for excitation; hence a more advanced approach is needed as a solution to make the signal viable for LIS.

4.13 Chapter conclusions

In this chapter, an analysis of the power optimizer as an excitation source was presented. A boost SMC was chosen over the buck and buck-boost configurations, because it allowed for the module's output current to remain continuous within the desired operational range – a crucial linearity property for implementing LIS. The analysis of the input current ripple showed that 50% duty ratio was a good operation point because in applications where the input current ripple needs to be eliminated for impedance measurements, a multiphase boost converter is the ideal excitation source. However, in applications where the input current is used as the excitation signal of choice, a single-phase boost converter is ideal as it provides the maximum current ripple at this operating point.

Modelling the dynamic system showed that when implementing impedance spectroscopy using the converter, the injection of a single duty cycle perturbation signal results in 2 outputs – an induced current at the input and an induced voltage at the output of the converter. The current loop analysis showed that the maximum attainable loop bandwidth was only 10% of the converter's switching speed. To ensure stability while performing impedance measurements, a lead compensator was designed to achieve such a bandwidth. The bandwidth of the voltage loop was limited by the RHP zero; however, because impedance measurements are performed at the input of the converter, this had negligible impact on impedance measurements.

The simulation results showed that under the designed control system, both the multiphase and single-phase converters achieved good transient and steady state responses – within the desired control and power specification. The faster transients of the multiphase showed larger output voltage overshoots than the single-phase converter – a finding that was noted for the experimental verification. The control also showed good setpoint tracking and disturbance rejection, at both the input and output; this is important when injecting perturbation signals at various operating points.

The chirp and PRBS signals showed good data acquisition times. However, both chirp and PRBS signals showed issues relating to excitation amplitude at varying frequencies. The major takeaway from the use of broadband excitation signals in converter-based impedance measurements is that the frequency responses of both converter and module AC EEC, affect the excitation amplitude and crest factor and hence, effective linear excitation frequency, of the induced excitation signal – thus the adoption of the signals through the SMC requires advanced implementation. The triangular wave, however, was of particular interest because it offered direct excitation through the use of the converter's own inductor current ripple, this improved the acquisition time for the signal significantly.

4.14 References

- [1] K. H. Kumar and G. V. S. K. Ra“, "A Review of Various DC-DC Converter Topologies for Photovoltaic Applicatio"s," 2021¹ 6th International Conference on Communication and Electronics Systems (ICCES), 2021, pp. 49-52, doi: 10.1109/ICCES51350.2021.9489070.
- [2] C. Deekshitha and K. L. Sheno“, "Design and simulation of synchronous buck converter for LED applicati"n," 2017² 2nd IEEE International Conference on Recent Trends in Electronics, Information & Communication Technology (RTEICT), 2017, pp. 142-146, doi: 10.1109/RTEICT.2017.8256574.
- [3] S. N. Soheli, G. Sarowar, M. A. Hoque and M. S. Hasa“, "Design and Analysis of a DC -DC Buck Boost Converter to Achieve High Efficiency and Low Voltage Gain by using Buck Boost Topology into Buck Topolo"y," 2018 International Conference on Advancement in Electrical and Electronic Engineering (ICAEEE), 2018, pp. 1-4, doi: 10.1109/ICAEEE.2018.8643001.
- [4] S. Sinha, Vinayak, A. Gullapalli, S. S. John and S. Kund“, "DC DC Boost Converter for Thermoelectric Energy Harvesti"g," 2018 International Conference on Current Trends towards Converging Technologies (ICCTCT), 2018, pp. 1-4, doi: 10.1109/ICCTCT.2018.8551039.
- [5] Ching-Tsai Pan, Shih-Kun Liang and Ching-Ming La“, "A zero input current ripple boost converter for fuel cell applications by using a mirror ripple circu"t," 2009 IEEE^E 6th International Power Electronics and Motion Control Conference, 2009, pp. 787-793, doi: 10.1109/IPEMC.2009.5157491.
- [6] A. K. Teeula and S. Kabol“, "A Reliable and Fast Response Buck Converter Based on Multiphase Convert"r," 2019²⁷th Iranian Conference on Electrical Engineering (ICEE), 2019, pp. 804-807, doi: 10.1109/IranianCEE.2019.8786512.
- [7] N. O. (. R. R. (. Peter Wellmann (Editor), Wide Bandgap Semiconductors for Power Electronics: Materials, Devices, Applications, 2 Volumes, Wiley.
- [8] R. Hoheisel, S. P. Philipps, F. Dimroth and A. W. Bet“, "Theoretical investigation on optimum bandgap energies of Ge-Based III-V Triple-Junction solar cells for long-term energy production on ma"s," 2009³⁴th IEEE Photovoltaic Specialists Conference (PVSC), 2009, pp. 001927-001930, doi: 10.1109/PVSC.2009.5411557.
- [9] S. Kazemini, K. Hadidi, A. Khoei and M. -N. Azarmanes“, "Effect of bandgap energy temperature dependence on thermal coefficient of bandgap reference volta"e," 2011²⁰th European Conference on Circuit Theory and Design (ECCTD), 2011, pp. 449-452, doi: 10.1109/ECCTD.2011.6043383.
- [10] J. Yu, Z. Li, Z. He, X. Jiang, C. Zhang and J. Wan“, "Performance Comparison of Traditional and JBS Integrated SiC MOSFETs in Si/SiC Hybrid Swit"h," 2019 IEEE Energy Conversion Congress and Exposition (ECCE), 2019, pp. 1918-1921, doi: 10.1109/ECCE.2019.8912713.

- [11] E. A. Jones, "Review and Characterization of Gallium Nitride," University of Tennessee, , Knoxville, 2016.
- [12] Martin Braae, "Control systems 1", University of Cape Town, Cape Town
- [13] Sang-Hoon Kim, "Control of direct current motors", in Electric Motor Control, 2017
- [14] S. Zeng, G. Zhou and S. Zho", "Small Signal Modeling and RHP Zero Analysis of Tri-state Boost Converter with Different Freewheeling Control Strategi"s," 2018 IEEE International Power Electronics and Application Conference and Exposition (PEAC), 2018, pp. 1-4, doi: 10.1109/PEAC.2018.8590582.
- [15] S. Kapat, A. Patra and S. Banerje", "RHP zero elimination with near optimum regulation in a current controlled tri-state boost converter through inductor current filteri"g," 2010 Power and Energy Conference At Illinois (PECI), 2010, pp. 83-87, doi: 10.1109/PECI.2010.5437153.
- [16] M. -F. Hung and K. -H. Tsen", "Study on the Corresponding Relationship Between Dynamics System and System Structural Configurations—Develop a Universal Analysis Method for Eliminating the RHP-Zeros of Syst"m," in IEEE Transactions on Industrial Electronics, vol. 65, no. 7, pp. 5774-5784, July 2018, doi: 10.1109/TIE.2017.2777386.
- [17] S. Moore and P. Barends", "Online condition monitoring of lithium-ion batteries using impedance spectroscopy," 2017 IEEE Energy Conversion Congress and Exposition (ECCE), 2017, pp. 5617-5624, doi: 10.1109/ECCE.2017.8096935.
- [18] Mark A. Haidekker, "Building blocks of linear systems" in Linear Feedback Controls (Second Edition), 2020
- [19] M. C. M. Teixeira, E. Assuncao and E. R. M. D. Machad", "A method for plotting the complementary root locus using the root-locus (positive gain) rul"s," in IEEE Transactions on Education, vol. 47, no. 3, pp. 405-409, Aug. 2004, doi: 10.1109/TE.2004.825068.
- [20] J. E. Salazar-Duque, E. I. Ortiz-Rivera and J. González-Llorent", "Modified perturb and observe MPPT algorithm based on a narrow set of initial conditio"s," 2016 IEEE ANDESCON, 2016, pp. 1-4, doi: 10.1109/ANDESCON.2016.7836255.
- [21] T. Latif and S. R. Hussai", "Design of a power optimizer based on SEPIC and buck topology using modified Incremental Conductance MP"T," 8th International Conference on Electrical and Computer Engineering, 2014, pp. 824-827, doi: 10.1109/ICECE.2014.7026999.
- [22] A. Fischer, A. Y. Kallel and O. Kanou", "Comparative Study of Excitation Signals for Microcontroller-based EIS Measurement on Li-Ion Batteri"s," 2021 International Workshop on Impedance Spectroscopy (IWIS), 2021, pp. 44-47, doi: 10.1109/IWIS54661.2021.9711770.
- [23] O. I. Olayiwola and P. S. Barends", "Characterization of Silicon-Based Photovoltaic Cells Using Broadband Impedance Spectroscopy," in IEEE Transactions on Industry Applications, vol. 54, no. 6, pp. 6309-6319, Nov.-Dec. 2018, doi: 10.1109/TIA.2018.2850025.

- [24] O. Alao and P. Barends", "Online Condition Monitoring of Sealed Lead Acid & Lithium Nickel-Cobalt-Manganese Oxide Batteries using Broadband Impedance Spectroscopy," 2018 IEEE Energy Conversion Congress and Exposition (ECCE), 2018, pp. 2026-2032, doi: 10.1109/ECCE.2018.8558352.

Chapter 5

Experimental considerations and impedance extraction system design for online implementation

5.1 System setup overview

The system was setup, as illustrated by Figure 5.1 and Figure 5.2. The system was designed such that the PV module charges a battery through an MPPT controlled converter – to emulate a small residential PV system with storage. The converter performs two tasks: firstly, harvesting maximum power from the module and regulating the power delivered to the battery, ensuring that the battery charges effectively; and secondly, performing online PV module BIS impedance extraction. The ambient temperature in the lab was carefully controlled to prevent both the module and prototype to overheat. A shell solar module was used to power the converter. A solar simulating light board was used as a solar simulator as shown in the figure. Two high-performance SiC MOSFET driver modules with integrated SiC MOSFETs were used in the power optimizer. Three auxiliary DC power sources were used to power the SiC MOSFET driver modules and the chips used in both the analog filters and PWM generator: 2 ESCORT EPS-3250 triple output DC power supplies with 3 channels with 2, 12V channels rated at 12W, and a single 5V channel rated at 10W; and a 60V 1200W QPX1200SP PowerFlex power supply. A RIGOL MS05072 digital oscilloscope was chosen because of its 70MHz 8Giga samples per second capability. This enables accurate monitoring of high frequency small amplitude signals, with minimal error. In offline implementation of both power regulation and impedance spectroscopy. The NI USB 6366 digital signal processing unit provides for both the power regulation and IS modulating signals. MATLAB & Simulink R2020a, was used as the main software interface for the processing unit.

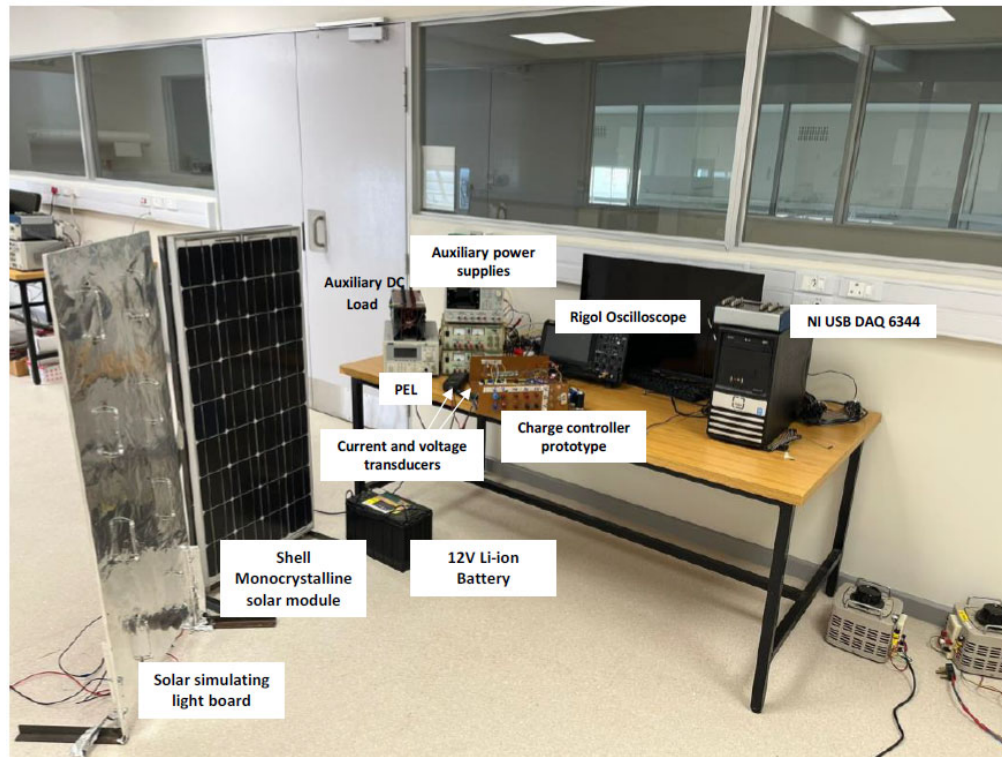


Figure 5.1 Online impedance extraction system laboratory setup

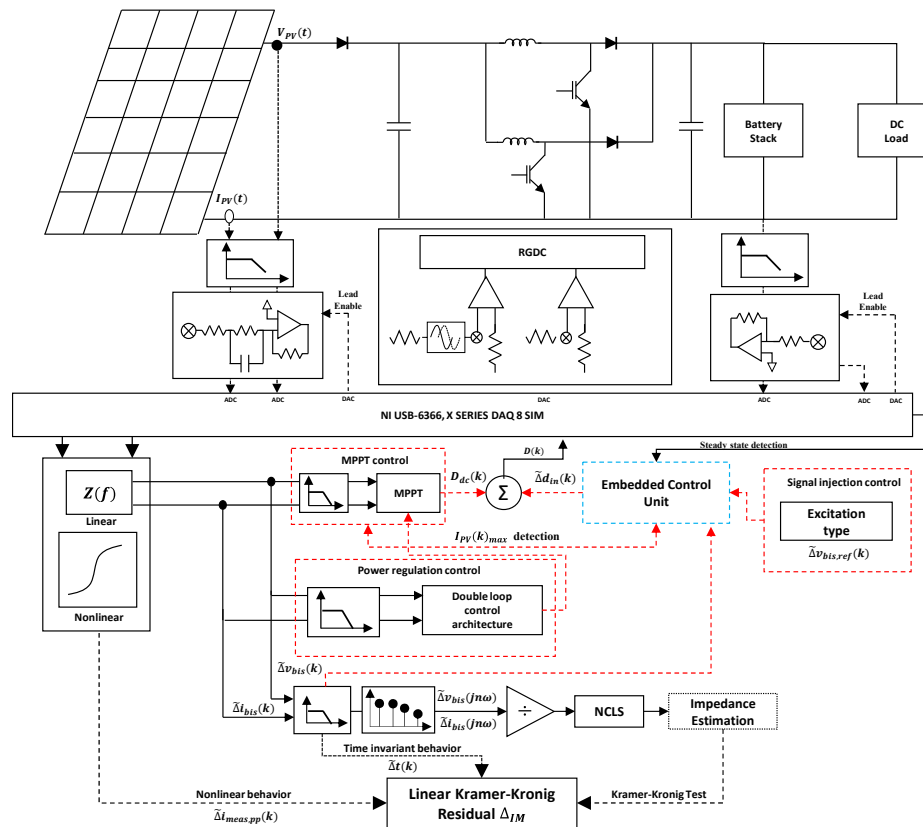


Figure 5.2 Online impedance extraction system setup block diagram

LabVIEW was also used as an interface for some aspects of both simulation and experimental analyses and results. Two high-bandwidth voltage and current transducers were used for processing the relevant control signals. An AUTOLAB FRA was setup with a 20A booster for higher current measurements. The ECU is the central processing unit for the power optimizer; it ensures that impedance measurements are performed in accordance with LTI requirements, while the power optimizer delivers regulated power to the battery. The algorithm for the ECU was designed within the NI DAQ and MATLAB's C-code.

5.2 Frequency Response Analyser

An AUTOLAB PGSTAT302N FRA was configured as a benchmark system to assess the validity of the impedance measurements and online condition monitoring system; the block diagram of this setup is illustrated in Figure 5.3 and actual setup. The base module 302N is implemented with a module FRA32M for electrochemical impedance measurements.

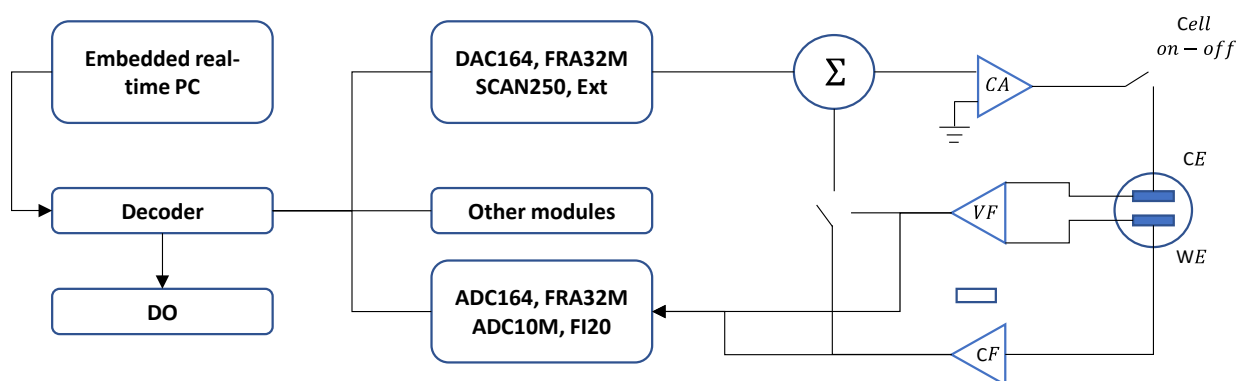


Figure 5.3 AUTOLAB PGSTAT302N frequency response analyser block diagram

Configured in galvanostatic mode, various small current AC perturbation single frequency sinusoids were added to the module's operating point and the voltage response of the module monitored. The device requires calibration to ensure the accuracy of the measurements. The calibration is required whenever the device operates in a new environment. Because of the increasing temperature when performing impedance measurements on the solar module, the calibration for 2 device capacitive parameters, C_1 and C_2 was performed after each set of impedance measurements.

The FRA module has a maximum compliance voltage of 30V with a potential range of $\pm 10V$. However, the module is only capable of providing a maximum perturbation amplitude of 500mV. This means that under laboratory conditions, the conditions were controlled such that the pseudolinearity requirement for solar module was met. The PGSTAT320N module has a measurement bandwidth of 1MHz – adequate for high frequency device excitation. The device is capable of 3 modes that make use of the bandwidth as follows. High stability mode is used for devices with impedance bandwidths $< 10kHz$; high-speed mode for devices with impedance

bandwidths $\in [10kHz, 125kHz]$; and ultra-high-speed mode for devices with bandwidth requirements $> 125kHz$. With the solar module chosen, high-speed mode was chosen, as will be discussed during the module's bandwidth characterization later in the chapter. Without the boost cell, the module is only capable of producing current response signals at a maximum of 2A, however, with the booster cell, the maximum response is 20A.

5.3 Power optimizer multiphase boost converter

A boost multiphase converter was designed in section 4.9.3. This section details the actual experimental components used and justifies their use in regulating power and impedance measurements.

5.3.1 *Prototype assembly design considerations*

The prototype was assembled and containerized to achieve a, $37cm \times 26cm$ footprint, at a height of just $8cm$ – sufficient for portable applications. The component selection for the prototype, considering experimental factors, is discussed in this section. In the introduction, it was mentioned that modifications to a multiphase converter would be made to allow for online impedance measurements; these modifications are summarized as follows:

- Silicon Carbide (SiC) MOSFET switches, wide bandgap semiconductors that have been shown to provide wide bandwidth and high switching capabilities in power converters, were used; this was to ensure a high switching frequency of $500kHz$, necessary for impedance extraction of a $100kHz$ impedance bandwidth PV module.
- A high performance SiC gate driver was used to improve the switching transients and reduce switching losses at high frequencies; this will improve power regulation efficiency while impedance measurements are being taken.
- The SiC MOSFET has short gate turn-on and turn-off times which ensures high switching capability, however, at high frequencies the switching losses increase; to solve this issue, an active soft switching cell was built to reduce the drain voltage and current time overlap. The cell was designed using a resonant inductor and capacitor aimed at ensuring resonance at the switching frequency.
- At high frequencies, however, the reverse recovery time of the switch requires reduction to minimize the losses; hence a fast recovery diode with a faster recovery time than the body diode of the high side SiC module was used. The diode bypassed the body diode of the resonant gate driver module to minimize the losses associated with the reverse voltage, especially at high operating duty cycles where the off time is significantly small.
- Lastly, a high frequency inductor was designed to reduce the losses associated with the skin effect from the high frequency small AC signal of the excitation signal.

i. Switch selection

As discussed throughout the study, wide bandgap semiconductor devices were chosen as primary switching devices for high frequency applications. As analysed in section 4.8, SiC MOSFETs were the most optimal choice for the application. Thus, C3M0120090J SiC MOSFETs were chosen for the prototype; these are N-channel semiconductors a 900V drain to source voltage breakdown capability – this is sufficiently larger than the voltage stress as designed for. It can handle 22A of continuous current from the drain to its source – again, a level that clears the maximum current stress the converter imposes on the switch. The ON-state drain to source resistance of the device is 170m Ω – a relatively low resistance compared to other semiconductor types. The rise time of the SiC MOSFET is 9ns; a turn-on delay of 12.5ns; a fall time of 5ns; and a turn-off delay of 15ns. This means that the minimum period that the MOSFET can handle for the switching converter is, 41.5ns – this is thus, a theoretical maximum switching frequency of 3MHz. With the converter switching at 500kHz – a period of 2 μ s – this will be sufficient in achieving good switching utilization.

ii. Inductor selection

A 33 μ H, Würth Elektronik 74436413300 leaded inductor was selected to meet the design specifications – with a $\pm$ 15% tolerance. The inductor was designed for 26 μ H, while the worst case for the chosen inductor is 28.05 μ H, which meets the inductance specification. The inductor is rated for a maximum DC current of 36A which is sufficiently larger than the expected system's maximum current at its peak, while the converter performs impedance estimation. More specifically, the converter can draw 4.76A at its maximum power point, with the broadband signal superimposed on this value, at a maximum of 10%, the system only reaches an approximate maximum current of 5.32A. The inductor has a specified DC resistance of 1.31m Ω – but a measured lab resistance of 70m Ω . At maximum system current, the inductor exhibits DC losses amounting to 372mW – this is only 0.4% of the system's rated power. The inductor has a maximum self-resonant frequency of 6.29MHz – this is the frequency at which the inductor's impedance approaches infinity – which is significantly lower than the module's 0.1MHz bandwidth and the converter's 0.5MHz switching speed. The inductor was thus, chosen as it met the minimum requirements for the design.

iii. Capacitor selections

A 1000 μ F Nichicon aluminium electrolytic capacitor was used for input voltage stabilization. With the module's maximum open circuit voltage of 21.8V, the capacitor is rated for a maximum voltage of 750V across its electrodes. The capacitor has a tolerance of $\pm$ 20% at 120Hz and 20°C. This is a low of 800 μ F for the capacitor at the specified frequency and temperature. Based on the desired value of 739 μ F, this value is sufficient.

A $470\mu F$ Nichicon aluminium electrolytic capacitor was used for the output. The capacitor is rated for a maximum DC voltage of $450V$. With the converter operating at maximum duty cycle, it reaches an approximate maximum output voltage of $87V$, which meets the capacitor's rated specs. The tolerance for the capacitor is the same as that for the input, thus resulting in a low capacitance of $376\mu F$ – sufficient for the application.

iv. ***DC cable selections***

The DC power cables used for the prototype were PVC insulated copper cables – chosen to achieve a good balance between current carrying capacity and system size, to avoid overdesigning. To determine the cable size, the IEC 60364-5-52 standard was used. The standard details the derating of DC cables under certain operational and installation condition. For the prototype, the cables would be on an insulated metal surface, which doesn't affect the cable derating as much as underground cabling. The major issue with the design is the derating due to the cables touching – this was done achieve a compact prototype. Because of the temperature control employed to test the solar module under various conditions, the derating temperature chosen for the system was chosen as $50^{\circ}C$. The cable was then selected by considering the maximum conducting for the system and applying the derating factors from the standard to obtain a worse case design – this is the cable size chosen such that when the system operates under these worse case conditions, the current carrying capability of the cable still meets the desired specifications without any danger to the circuit.

v. ***Diode selection***

A high frequency diode was also used to minimise the reverse recovery losses of the SiC switch's body diode at high frequencies at the highest duty cycle of 83%, the off-time for the converter at this operating point is $320ns$; that means that the recovery time for each of the diodes must be less than this value. A RHRP30120 fast recovery diode was used. The diode is rated for a recovery time of $85ns$ at $30A$; which is adequate for the application. Because of its power characteristics, this diode was subsequently used as the power stage diode.

5.3.2 Power MOSFET Driver

A power driver circuit is required to ensure that the power MOSFETs have a sufficient voltage at their gate to fully switch ON. A CRD-5FF0912P, SiC MOSFET High-Frequency Evaluation Board for 7L D2PAK with two C3M0120090J SiC MOSFETs installed, shown by Figure 5.4, was used as a power driver. The evaluation board was designed to demonstrate high switching performance.

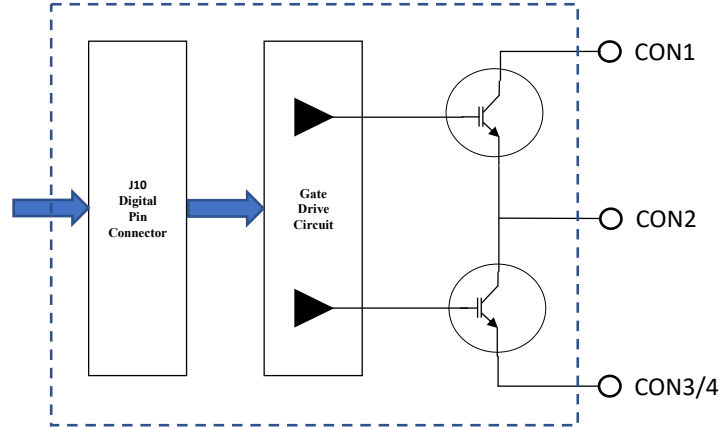


Figure 5.4 SiC MOSFET high-frequency evaluation board block diagram

Each SiC MOSFET has a gate driver circuit which consists of a 2A isolated gate driver chip and an isolated regulator converter, rated at $2W$, that enables the MOSFET to operate at a maximum of $3MHz$. The chip of the gate driver can provide up to $1200V$ of isolation between the low and the high voltage drives. The regulators are powered with $12V$ at their inputs. The evaluation board also contains a series of jumpers that enable the gate drivers to be configured as a low-cost unipolar i.e., $0V$ to $15V$ gate driver or as a high-performance i.e., $-3V$ to $15V$, bipolar gate driver.

5.3.3 Wide bandgap SiC semiconductor module configuration for high-speed performance

For effective impedance extraction, the multiphase boost converter was designed to switch at $500kHz$ using capable wide bandgap SiC modules. The module driver was then configured for high performance bipolar switching as shown in Figure 5.5. This is because when driving the high powered SiC MOSFET at high frequencies, there is a risk of a Miller effect induced at switch-on if a large voltage differential appears across the power module.

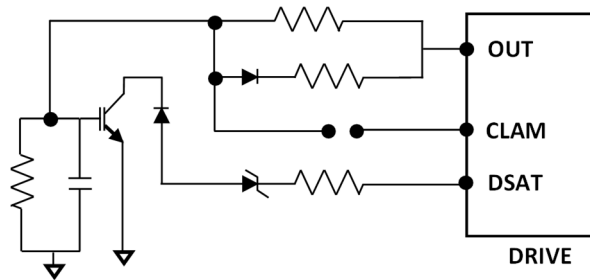


Figure 5.5 Silicon Carbide MOSFET module configuration for high performance bipolar switching with Miller clamp

When this happens, current would flow into the gate of the SiC module through one of the SiC gate capacitances. If this voltage is large enough such that the gate voltage exceeds the semiconductor's threshold voltage, the device would operate at a lower efficiency due to parasitic switch-on – and eventually device failure. Bipolar switching mitigated this, by driving the SiC gate ($-3V$) negative with respect to the source such that, when a

current spike occurs, the gate voltage is below a set common threshold. The gate driver consists of a 2A isolated gate driver chip with a 2W isolated DC-DC converter to enable high performance switching.

5.3.4 Active soft switching cell design

To better manage switching losses, at the 500kHz switching speed for the multiphase converter, a zero-voltage active soft switching cell was designed and assembled. The cell was designed to achieved zero-voltage, as well as zero-current switching, during the switching transients. The cell's structure is illustrated in Figure 5.6; from the structure, inductor L_s and capacitor, C_s were designed to required resonance.

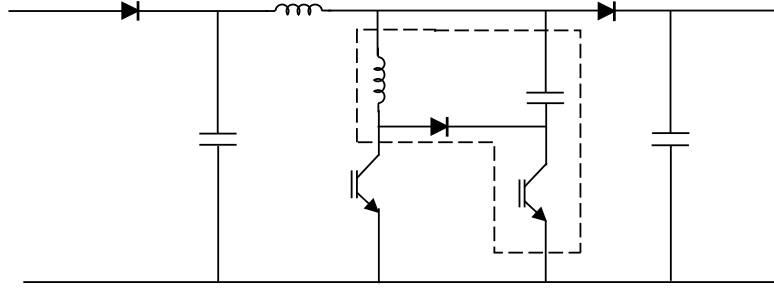


Figure 5.6 Boost converter per-phase active soft switching cell

To determine the inductance, the inductor current during the converter's OFF state can be characterized as:

$$i_{L_s}(t) = i_{L_s}(t_i) + \frac{1}{L_s} \int_{t_i}^t V_{bat} dt \quad (5.1)$$

$$= i_{L_s}(t_i) + \frac{V_{bat}}{L_s} (t - t_i)$$

Considering in initial state where the inductor current is zero, the resonant time constant is:

$$(t - t_i) = \frac{I_{PV} L_s}{V_{bat}} \quad (5.2)$$

When considering the fact this time constant is approximately a 5th of the resonant period, the resonant inductor works out to 1.15μH. To ensure resonance, the energy stored in the cell's electric field should be significantly larger than the energy stored in the cell's magnetic field; that is, the stored energy in the resonant capacitor must be greater than the stored energy in the resonant inductor, such that:

$$\frac{1}{2}C_s V_{bat}^2 \geq \frac{1}{2}L_s I_{PV}^2 \quad (5.3)$$

The inequality described in (5.3) requires a minimum capacitance to ensure that there is sufficient energy in the capacitor. Solving the equation, yields a minimum capacitance value of $89nF$.

5.4 Power optimizer data acquisition and generation system

The system setup for data acquisition and generation is illustrated by the block diagram. As illustrated by Figure 5.7, the transducers capture either the excitation or control signal.

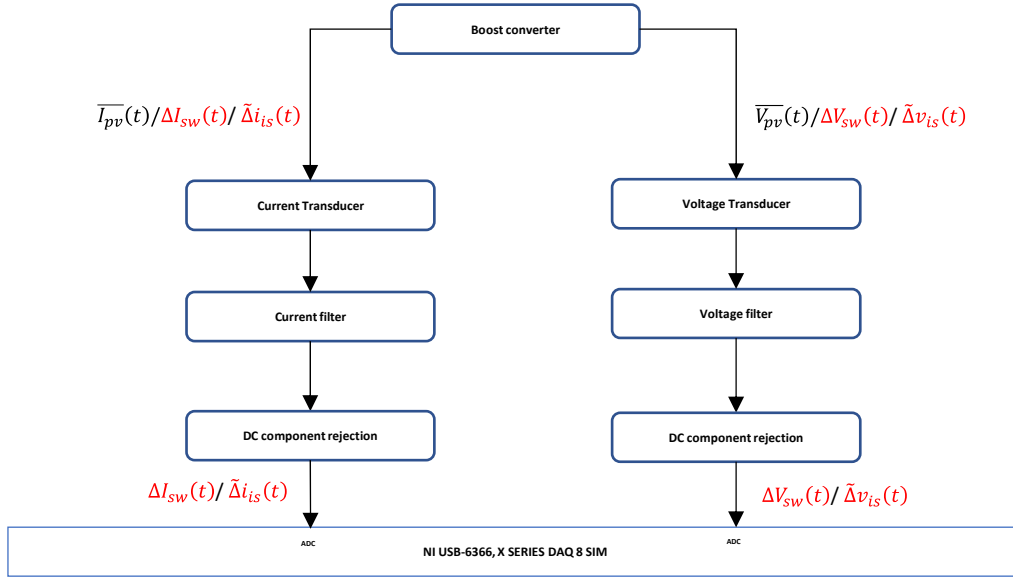


Figure 5.7 Data acquisition system setup block diagram

The transducers convert the voltage and current signals into compound DC electrical signals with varying amplitudes. Because of the wide bandwidth the transducers have, significant noise is picked up and amplified by the devices; low pass filters are thus designed to suppress this high frequency noise. To eliminate the DC component, $\overline{V_{dc}}(t)$ from the excitation signal, a DAC based method is used instead of a high pass filter, for reasons to be elaborated upon. The digital signal processing unit thus processes signals with less interference from noise and the phase and amplitude errors associated with it.

5.4.1 Transducers

For voltage data acquisition, a Tektronix P5200A high-voltage differential probe was chosen. The probe has a DC gain of about $80dB$, and a $60dB$ gain at a bandwidth of $100kHz$. This translates to a gain of 1000 at the maximum measured frequency of the PV module. The probe can measure $130V$ with a voltage resolution of $100mV$. With the module operating at 75% of its maximum power, as discussed earlier, a perturbation amplitude of $1.23V$ and a gain of 1000, this translates to 12,300 distinct voltage levels for the voltage excitation signal – a good voltage resolution for data acquisition.

A 60A Pintec PT-350 high frequency current probe was used for current data acquisition. The probe has a maximum bandwidth of $50MHz$ which is sufficient in tracking the module's impedance with minimal error. It has an accuracy of 1% which is approximately, $\pm 5mV$. The probe can acquire a minimum current value of $20mA$. The PV module sees its lowest impedances at under high irradiance conditions, low temperature, and higher frequencies. For an $80W$ monocrystalline module, the typical impedance absolute magnitude, under such conditions is approximately, 2Ω – when measured using an FRA. With a perturbation amplitude of $1.61V$, as discussed earlier, a response current with an amplitude of $800mA$. The probe's resolution can sufficiently reconstruct the response signal.

5.4.2 Analog filters

As discussed earlier, analog filters were designed to eliminate high frequency noise and provide stability for both the excitation and response signals. The details are discussed in this section.

i. Filter design

Sallen-Key second order active low pass filters were chosen and designed for the transducers. This configuration was chosen because of several advantages this type of filter has, these are: simplicity in the design, easy cascade of first and second order designs, and steep roll-off for second order configuration; this configuration is illustrated in Figure 5.8. The other advantage of this configuration is that the high input impedance of the OP-Amp prevents the filter output from overloading. The major drawback of the Sallen-key filter is that the voltage gain, and the magnification factor are similar because of the use of an OP-Amp. Because use of the non-inverting input, the gain must be kept below 3, otherwise the filter may become stable.

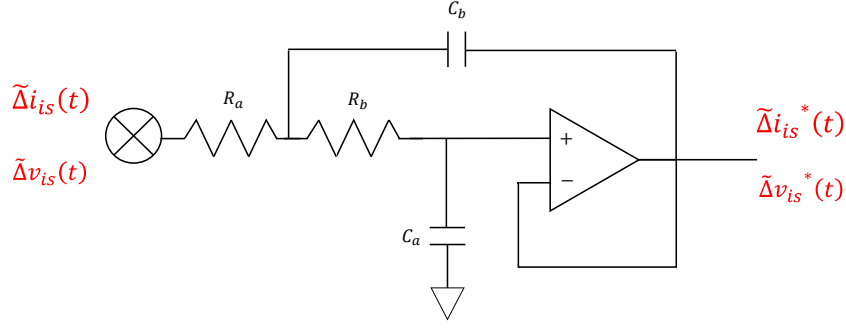


Figure 5.8 Sallen-key second order low pass filter configuration

At low frequencies, capacitors C_a and C_b appear as open circuits, so the input signal is passed resulting in an output. At higher frequencies, C_a and C_b appear to the sinusoidal input signal as short circuits, so the signal is passed to ground. At the cut-off frequency, the impedance of C_a and C_b will be the same value as R_a and R_b , so the positive feedback produced through C_b provides voltage gain and increase in output signal magnification, Q . The transfer function of the filter is:

$$\frac{\tilde{\Delta}v_{is}^*(s)}{\tilde{\Delta}v_{is}(s)} = \frac{(R_a C_a R_b C_b)^{-1}}{s^2 + ((R_a C_b)^{-1} + (R_b C_a)^{-1})s + (R_a C_a R_b C_b)^{-1}} \quad (5.4)$$

The specifications for the filters were resolved from the requirements to perform impedance spectroscopy. The filter is to be designed to achieve a low frequency gain of 2, with an input impedance, R_a of $10k\Omega$. High gains in the amplifier aren't required because the transducers used can achieve high bandpass gains. The filter should achieve a passband with a cut-off frequency equivalent to the bandwidth of the DUT – in this case, $100kHz$. With these specifications, R_b , C_a and, C_b were determined to be: $10k$, $1nF$ and $2.5nF$, respectively. This led to a cut-off frequency of $100.66kHz$ – close to the desired value.

ii. **Impedance measurement phase error**

As discussed earlier, both the voltage and current probes' high bandwidth enables them to introduce negligible phase error. Both the voltage and current transducers' high bandwidth lead to high frequency noise being detected. This is particularly problematic because the ADC of the data acquisition system introduces sampling phase error between the measured signal and the output signal with increasing frequency. The excitation and response signals are sampled serially such that, with the ADC of the DAQ configured to a sampling frequency of $550kHz$, the delay between two consecutive samples is, $2\mu s$ – a, $9.57 \times 10^{-3} rad.s^{-1}$ phase error at the cut-off frequency – which increases with system noise. To reduce the high frequency noise and hence, phase error, second order low passing, Sallen-Key analog active filters, were designed. Active filters were chosen such that its

active gain increases the magnitude of both excitation and response signal for better SNR during processing. The phase response of the filter is:

$$\angle \frac{\tilde{\Delta}v_{is}^*(s)}{\tilde{\Delta}v_{is}(s)} = -\tan^{-1} \left(\frac{\omega[R_a C_b + R_b C_a]}{1 - \omega^2[R_a C_a R_b C_b]} \right) \quad (5.5)$$

The filters were designed such that the, $-3dB$ bandwidth is, $50kHz$ and a gain of 10. Equation (5.5) evaluates to a maximum phase of, $6.66 \times 10^{-2} rad. s^{-1}$ at the cut-off frequency. As discussed in the section IV, capacitances have variance dependent on their tolerance – which is, 1%, while the resistors used in the design have a combined tolerance of 3%. This is a phase of between, $[6.36; 6.92] \times 10^{-2} rad. s^{-1}$. The range of this phase is the maximum possible phase error between the excitation and response signals at the cut-off frequency due to the filter – which is $5.60 \times 10^{-3} rad. s^{-1}$. The total expected phase error is therefore, $1.51 \times 10^{-2} rad. s^{-1}$ at $50kHz$.

5.4.3 Digital signal processing unit

As mentioned in the introduction, an NI USB 6366 was setup for data acquisition and generation. It is a 16bit device, capable of $2MSs^{-1}$ per channel. The device has a maximum clock frequency of $10MHz$ – considered a medium power processor. This implies at most, 18 samples per period for the module's excitation and response signals at the largest frequency. It has 2AO, each with $3.33 MS/s$. It consists of 24 digital input-output pins, serving as multifunction I/O Device. It offers analog I/O, digital I/O, and four 32-bit timers for PWM and frequency. The most important feature of the device is that it offers simultaneous parallel sampling which is vital for measuring the impedance from an excitation signal measured from one and a response signal measured in another. This eliminates phase related issues that arise from series sampling. Its onboard NI-STC3 timing and synchronization technology delivers advanced timing functionality, including independent analog and digital timing engines and refrigerable measurement tasks. The NI-DAQ ADC can convert the excitation voltage into a 16 bit digital value. The ADC has a maximum range of $[-10; 10]V$ and hence, a maximum ADC resolution of, $(20/2^{16} = 0.31mV)$. For the maximum perturbation amplitude of $1.23V$, this results in 3967 possible distinct voltage levels – sufficient in reconstructing the excitation signal. For the current response signal, the lowest amplitude observed was, $620mA$ at, $108kHz$ – resulting in 2000 possible voltage levels. The low number of samples at this frequency leads to the largest quantization error for the excitation and response signals. The quantization error is the difference between the analog signal and the closest available digital value at each sampling instant from the signal processing unit's ADC. This error also introduces quantization noise, to the sampled excitation signal. For an ideal ADC, the signal to noise ratio can be expressed as a relationship between the maximum ADC resolution and quantization noise, which is, $(-20\log 1/2^n)$; where, n is the resolution of the ADC.

5.4.4 DC component rejection

To eliminate the DC components, $V_{dc}(t)$ and $I_{dc}(t)$ from the excitation and response signals, a DAC based method is used instead of a high pass filter. The lower frequency of the PV module is, 10Hz , which will require a low frequency cut-off for the high pass filter that is at most, 5Hz . This will lead to a longer filter settling time which will result in an increase in the data acquisition time. This is not desirable as the condition in which PV modules operate, change rapidly – hence, a faster data acquisition time is preferred.

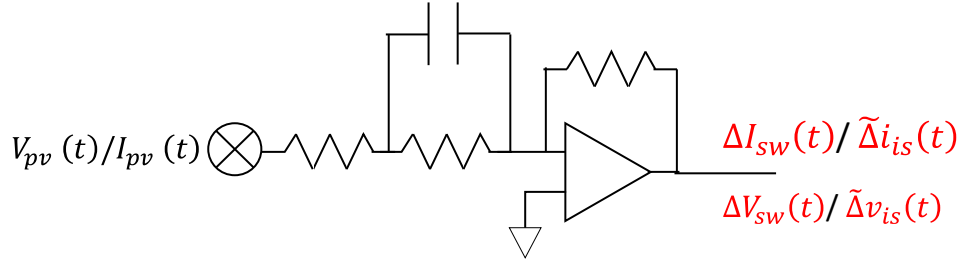


Figure 5.9 DC component rejection control technique

As shown by Figure 5.9, the DAC output is subtracted from the broadband voltage excitation signal by a differential amplifier. Feedback closed loop with lead compensation was used to eliminate the DC component as shown. The lead compensator ensures a high loop bandwidth such that the DAC output can accurately follow the output of the PV module. A low pass filter is used to eliminate any high frequency noise from the measured signal.

5.4.5 System's linearity during impedance measurements

To assess the linearity of the system while the measurements are taken, the following conditions and assumptions were set: the module is operating under low-light conditions, at $250\text{W}/\text{m}^2$; and the increase in the module's current and voltage due to the response signal is negligible due to the lower response current magnitude under low irradiance conditions. For impedance measurements taken while the panel operates at maximum power, the maximum perturbation amplitude must be, 1.61V . This increases the module's operating voltage to a peak of 18.75V and a low of 15.53V ; and the operating current thus, decreases to a low of 0.99A , and high of 1.26A . At this operating point:

$$\left. \frac{dI_{p-}}{dV_{p+}} \right|_{P_{max}} = -0.13$$

While:

$$\left. \frac{dI_{p+}}{dV_{p-}} \right|_{P_{max}} = -0.037$$

Thus, the net rate at which the module's current changes due to the positive peak of the voltage excitation signal, evaluated at the maximum power point, is more than 3 times than that from the negative peak. The lack of collinearity in these points, thus implies that the measurements would be taken in a non-linear range of operation, and hence, producing invalid impedance results. This problem is particularly significant at higher frequencies because the module's impedance is at its lowest. The response current at such frequencies would, thus have a larger amplitude compared to lower frequencies. The consequence of this is a variable non-linear range of operation for all excitation frequencies.

As such, operating voltage was pulled down to, $6.8009V$ and thus increasing the module's current to, $1.2811A$. This voltage is still sufficient in charging the battery, but at a slower rate compared to its capacity. Based on the THD observed for the BIS current response, the maximum perturbation amplitude of the induced voltage excitation signal, $\tilde{v}_{bis}(t)$ was, $1.13V$, resulting in the module's operating voltage to a peak of, $5.67V$ and a low of, $7.93V$; and the operating current decreases to a low of, $1.2753A$, and high of, $1.2861A$.

$$\left. \frac{dI_{p-}}{dV_{p+}} \right|_{40\%P_{max}} = -0.0047$$

While:

$$\left. \frac{dI_{p+}}{dV_{p-}} \right|_{40\%P_{max}} = -0.0041$$

These points show collinearity and thus, implying system linearity while impedance measurements are being taken. Under-high irradiance conditions, similar linearity characteristics can be expected with the response current at lower frequencies. At higher frequencies however, the magnitude of the response current is largest and thus, has a larger impact on the average gradient during each period, which means that a peak response predictor variable in a regression model cannot be linearly predicted from the others with a substantial degree of accuracy, thus increasing the uncertainty related to the linearity of the system if it operates at maximum power. Hence, although the solar module can be designed to operate at maximum power during the measurements, the uncertainty around the system's linearity at that point is significant, as this point is the boundary between the module's voltage and current source regions. As such, when taking measurements in the current source region, before impedance measurements take place, the control unit sends a signal that momentarily reduces the module's operating voltage, to points around $40\% V_{max}$ – thus increasing the operating current. This will ensure that the module is operating purely as a current source during the impedance measurements – such that the maximum allowable system duty ratio is not exceeded while maintaining the battery's charging voltage. The same logic can be used for measurements in the module's voltage source region. Hence, the module's current or voltage are reduced or increased by $\pm 40\% V_{max}$ by approximately to ensure linearity. Because the signals undergo signal processing, the

excitation signal amplitudes are required to be as large as possible without pushing the signal to a non-linear range. This is because a lower signal amplitude, results in a lower SNR. Hence, the SNR of the current response signal was observed for various voltage perturbations to determine lowest perturbation amplitude. It is worth noting that the % V_{max} operating point for signal injection, is not set at 40% for all scenarios, but rather chosen based on power characteristics of the module at a particular operating point.

5.4.1 Quantization error analysis

In simulation implementation, the voltage excitation and current response signals have infinite resolution, and can take on any value; however, in experimental implementation, the DAQ's bidirectional ADC quantizes the signals into a set of discrete samples. The error between the sampled and actual signals is the quantization error. To analyze this error, let us consider the least significant bit of the ADC, δ such that the quantizer's characteristics are:

$$Y_{\delta}(t) = \lambda\delta \mid Y(t) \in \left[\lambda\delta - \frac{\delta}{2}; \lambda\delta + \frac{\delta}{2} \right] \quad (5.6)$$

It should be noted that the output of the quantizer, $Y_{\delta}(t)$ has half and quarter-wave symmetries, thus the input signal, $Y(t)$ and quantizer's output have the same zero-crossings at every interval. This means that the quantization process will introduce negligible phase error to the impedance measurements. There will, however, be an error in the magnitude of the quantized signal. Consider an integer, Λ , such that:

$$\delta \left[\Lambda - \frac{1}{2} \right] \leq \max_{\delta} Y(t) \leq \delta \left[\Lambda + \frac{1}{2} \right] \quad (5.7)$$

Thus, for sampling impedance perturbation and excitation signals, there can only δ quantization levels for the approximation of the signals' peak response, $\max_{\delta} Y(t)$. However, because, as discussed earlier in the chapter, the ADC has a maximum of 2^{n-1} quantization levels, there is an underutilization of the ADC which causes the aforementioned magnitude error. To mathematically quantify the quantized signal at quarter-wave symmetry, its Fourier series was analyzed as follows:

$$\Gamma_{\delta}(t) = \frac{8}{T_n} \int_0^{\frac{T_n}{4}} Y_{\delta}(t) \sin(\omega_n t) dt \quad (5.8)$$

Which evaluates to:

$$\Gamma_{\delta}(t) = \frac{8}{T_n} \left[\sum_{\lambda=1}^{\Lambda-1} \left[\int_{t_{\lambda}}^{t_{\lambda+1}} \lambda \delta \sin(\omega_n t) dt \right] + \int_{t_{\Lambda}}^{\frac{T_n}{4}} \delta \Lambda \sin(\omega_n t) dt \right] \quad (5.9)$$

Now for an n bit ADC with an intersection with the, i^{th} quantization level at time, t_i , the signal can be characterized as:

$$\max_{\delta} Y(t) \sin(\omega_n t_{\lambda}) = \delta \left[\Lambda - \frac{1}{2} \right] \quad (5.10)$$

Where:

$$\omega_n t_{\lambda} = \sin^{-1} \left[\frac{\delta}{\max_{\delta} Y(t)} \left(\Lambda - \frac{1}{2} \right) \right]$$

$$\cos(\omega_n t_{\lambda}) = \sqrt{1 - \left[\frac{\delta}{\max_{\delta} Y(t)} \left(\Lambda - \frac{1}{2} \right) \right]^2}^{-1}$$

Substituting result yields:

$$\Gamma_{\delta}(t) = \frac{8\delta}{2\pi} \left[\sum_{\lambda=1}^{\Lambda-1} \lambda \left(\sqrt{1 - \left(\frac{\delta \left(\lambda - \frac{1}{2} \right)}{\max_{\delta} Y(t)} \right)^2} - \sqrt{1 - \left(\frac{\delta \left(\lambda + \frac{1}{2} \right)}{\max_{\delta} Y(t)} \right)^2} + \Lambda \sqrt{1 - \left(\frac{\delta \left(\lambda - \frac{1}{2} \right)}{\max_{\delta} Y(t)} \right)^2} \right) \right] \quad (5.11)$$

The error is worst when the maximum amplitude is such that: $\max_{\delta} Y(t) = \delta(\Lambda - 1/2)$:

$$\Gamma_{\delta}(t) = \frac{8 \max_{\delta} Y(t)}{2\pi(\Lambda - 1/2)} \left[\sum_{\lambda=1}^{\Lambda-1} \lambda \left(\sqrt{1 - \left(\frac{\delta \left(\lambda - \frac{1}{2} \right)}{\left(\Lambda - \frac{1}{2} \right)} \right)^2} - \sqrt{1 - \left(\frac{\delta \left(\lambda + \frac{1}{2} \right)}{\left(\Lambda - \frac{1}{2} \right)} \right)^2} \right) \right] \quad (5.12)$$

Thus, the percentage error is the amplitude of the quantized signal can be characterized as:

$$\Upsilon_e(t) = 100 \left[1 - \frac{4}{\pi(\Lambda - 1/2)} \left[\sum_{\lambda=1}^{\Lambda-1} \lambda \left(\sqrt{1 - \left(\frac{\delta \left(\lambda - \frac{1}{2} \right)}{\left(\Lambda - \frac{1}{2} \right)} \right)^2} - \sqrt{1 - \left(\frac{\delta \left(\lambda + \frac{1}{2} \right)}{\left(\Lambda - \frac{1}{2} \right)} \right)^2} \right) \right] \right] \quad (5.13)$$

This error percentage gives a good indication of the expected error is the quantized signal, as will be discussed during the analysis of the generation of perturbation signals.

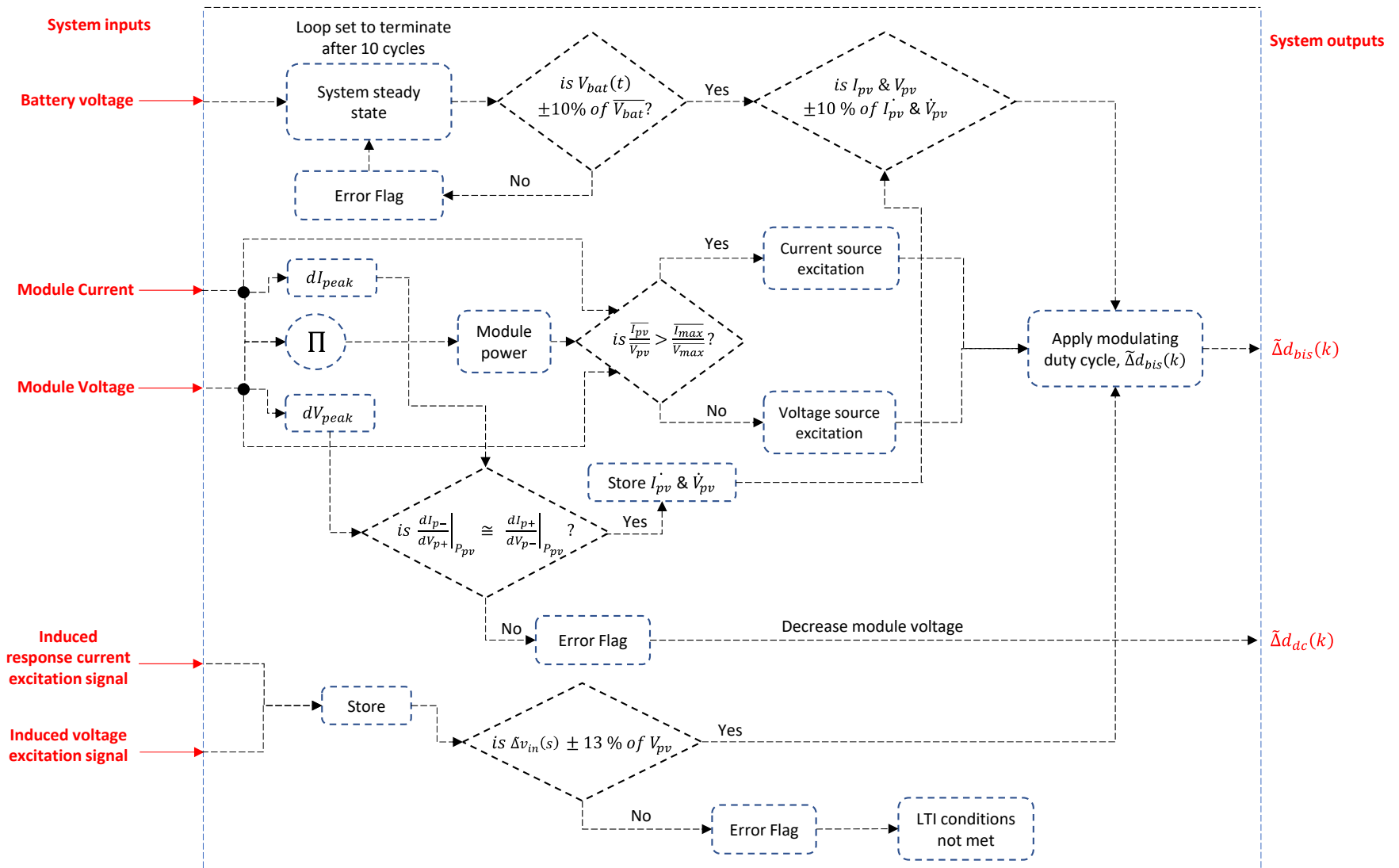
5.5 The embedded control unit algorithm design

As mentioned earlier the ECU is responsible for processing the system's impedance extraction algorithm – it is illustrated in Figure 5.10. The ECU first ensures that impedance extraction only takes place once the power regulation system has reached steady state; it does this by monitoring the change in the voltage delivered to the batteries in real-time. The change must be within 10% of the average value for the system to be considered in steady state. Since the BIS measurements are to be taken while the PV module is operating either as a current source or voltage source, incremental conduction MPPT control algorithm was chosen. This is because the PV module operates as a current source for all currents that are larger than the current at maximum power, while operating as a voltage source for all currents lower than the current at maximum power; only MPPT can track the maximum current point. The ECU thus, lowers or increases the PV module's operating voltage before impedance extraction takes place – this increases or decreases the module's operating current, depending on the desired excitation region.

At these points, the module would operate either as a current or voltage source as it provides a steady current for various operating voltages. Before impedance estimation commences, the control unit notes the PV module's operating current and voltage. Once impedance extraction has ceased, the control unit notes the module's current and voltage once again. It then compares the values to ensure that the change in voltage and current is less than the peak amplitude of the excitation signal required to keep the system linear. The control unit also notes the peak PV module power response such that:

$$\left. \frac{dI_{p-}}{dV_{p+}} \right|_{P_{pv}} \cong \left. \frac{dI_{p+}}{dV_{p-}} \right|_{P_{pv}} \quad (5.14)$$

Impedance estimation results that do not satisfy the equation for all intervals within the signal acquisition time, are then rejected by the control unit as they were taken in a non-linear range. To ensure that the AC components of the PV module's voltage and current are extracted accurately without the DC component, the ECU was designed such that it has an embedded feedback control system with lead compensation. The lead compensator ensures a high bandwidth for the measured signals. All the photos of the actual setup can be found in the appendices.



5.6 Experimental results and discussions

These tests were performed as they were deemed necessary for the realization of online linear impedance spectroscopy. The multiphase configuration was used as the primary converter, unless stated otherwise. The converter's power stage was first tested under various conditions that could be encountered by the system; the system's maximum power tracking specification was then tested; the optimal module perturbation range was analyzed using the FRA; the I/V curve characteristics were assessed and both series and shunt resistances were characterized; and lastly, the FRA was used to characterize the module under various operational conditions.

5.6.1 *Solar simulating light board*

As mentioned in the introduction, the tests were performed in a laboratory setting using a solar simulating light board; this was done to have better management of the system's-controlled variables. Eight DE 118mm, ECO 120V, 500W R7S SV1 SYLVANIA, linear double ended light bulbs were used in assembling a solar simulating light board for the module. The bulbs have a 3000K colour temperature. Although this lies near the infrared spectrum, they were chosen to achieve good temperature experimentation. The bulbs, assembled onto the light board, can increase the temperature of module as desired depending on the distance between the board and module, without significantly affecting the illumination on the module, this is because the bulbs have a luminous flux rating of 8750lm. Both the light board and module have the same dimensions of $1482 \pm 1mm \times 680 \pm 1mm$; this ensures a high concentration of light with minimal possibility of shading spots. Depending on the perpendicular distance between the light board and solar module, there will be various levels of insolation and temperature on the module. This gives good control of the module's input parameters. Clearly, the closer the module is to the light board, the higher the insolation and thus, higher temperature. For test that require higher insolation but lower temperature, a blower fan was used reduce the temperature to a lower steady state value. There are 2 graphs of concern regarding the light board, namely, the variation of the irradiance with perpendicular distance at a constant temperature, and the variation of the temperature at constant irradiance. To better emulate conditions in the field under a typical summer's day, a nominal irradiance of, $800 W/m^2$ and a nominal temperature of, $30^\circ C$ was chosen for the measurements, unless stated otherwise. The irradiance was measured using an irradiance meter, while the temperature was measured using a temperature gun. Figure 5.11 shows that, while both irradiance and temperature on the module decrease with increasing distance, the irradiance decreases geometrically, while the temperature decreases linearly at lower perpendicular distances. For the irradiance, a perpendicular distance of approximately 12cm was required to maintain the nominal $800W/m^2$. While for the temperature, a blower fan was required to cool the module to 30 degrees and below.

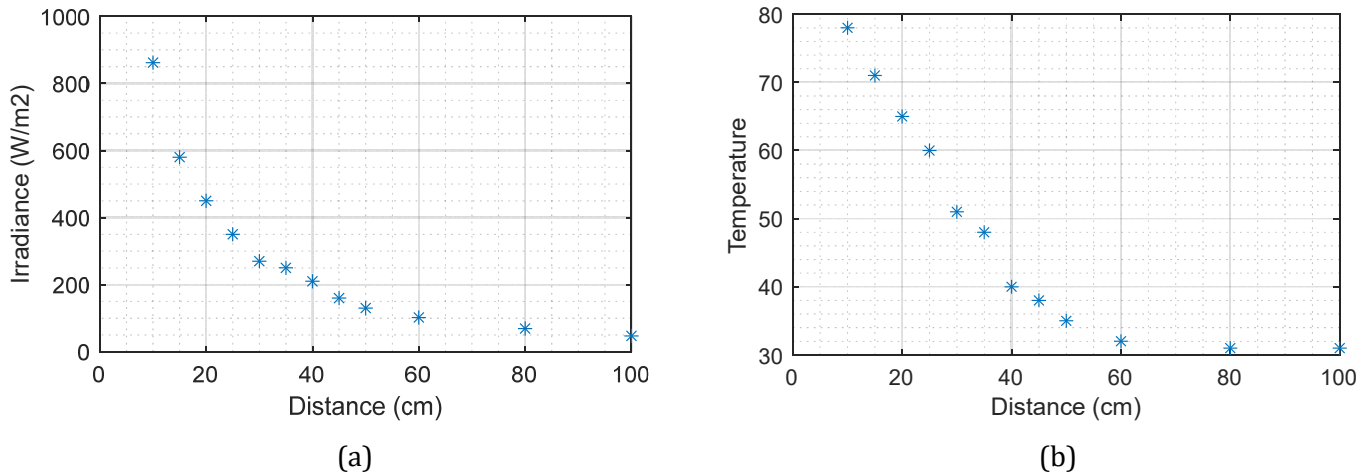


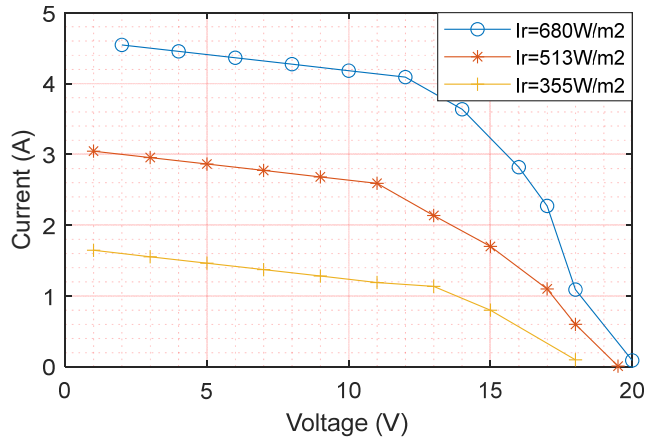
Figure 5.11 Solar simulating light board performance with perpendicular distance (a) irradiance (b) temperature

5.6.2 Solar module I/V curve characteristics

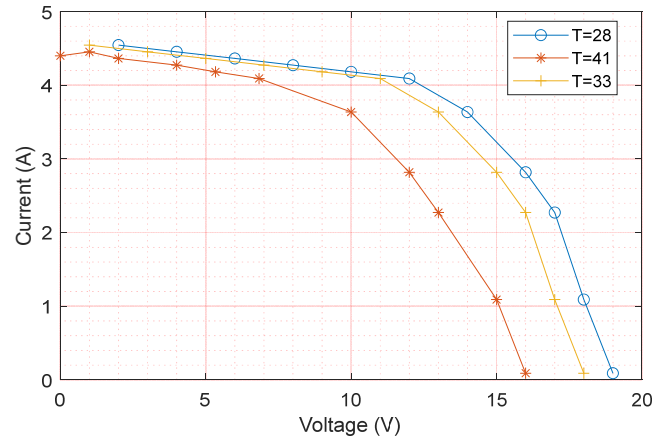
The module's AC parameters were tested based on the IEC-61853-1 standard. The objective of the standard is to, firstly, define a testing and rating system, which provides the PV module power (watts) at maximum power operation for a set of defined conditions; and secondly, provide a full set of characterization parameters for the module under various values of irradiance and temperature. The standard requires the performance evaluation of the module through a 23-element maximum power, P_{max} matrix at 4 temperature levels and 7 irradiance levels, as shown in Table 5.1. The matrix was generated using the indoor laboratory solar simulator such that the impedance parameters were determined for each element in the matrix. Due to the limitations in the light board and set up, irradiances above $800\text{W}/\text{m}^2$ proved to be unattainable. Because to achieve high irradiances, the distance between the board and solar module must be small, temperatures below 25°C were difficult to achieve. Power values at high temperature and low insolation were also difficult to achieve as the module and light board were too far apart. The module's I/V curve characteristics were therefore, tested based on the power matrix. An LD400P 400W 80V electronic load was employed to emulate a linear load to produce the I/V curve of the module under laboratory conditions.

Table 5.1 Solar module test power matrix

Irradiance (W/m^2)	Temperature ($^\circ\text{C}$)							
	28		30		35		40	
	Power (W)	%eff	Power (W)	%eff	Power (W)	%eff	Power (W)	%eff
50	3.38	5.41%	3.38	5.41%	N/A	N/A	N/A	N/A
100	7.32	5.86%	7.04	5.63%	N/A	N/A	N/A	N/A
200	N/A	N/A	14.36	5.74%	12.95	5.18%	N/A	N/A
400	N/A	N/A	29.57	5.91%	26.47	5.29%	N/A	N/A
600	N/A	N/A	N/A	N/A	40.27	5.37%	35.76	4.77%
700	N/A	N/A	N/A	N/A	54.06	5.41%	47.87	4.79%



(a)



(b)

Figure 5.12 Module I/V curve characteristics of the monocrystalline module (a) varying irradiance (b) varying temperature

5.6.3 DC-DC power stage closed-loop results

The power stage results analysed were the steady state input and output voltage results, as the module was operated in voltage control mode. The system ability to track various voltage step commands was assessed. The load step – both step-down and step-up – was assessed. Lastly, the ripple at both input and output of the converter was analysed for the converter types. The results were performed using the multiphase converter, unless stated otherwise.

i. Process control experimental design considerations

The control system, as defined by (4.72) and (4.73), was discretized using the backward difference equation such that, the "s" plain can be mapped to "z" plain through the equation, $s = z - 1(Tz)^{-1}$; where, T is the control sample time chosen to be, 2×10^{-6} , as multiples of a fixed step chosen for MATLAB's solver configuration. The same specifications used for the simulation results were adopted, except that the settling time specification of the system was adjusted to be, 300ms to accommodate the transients of the practical system. This meant that the gain of the system was reduced dragging the closed poles along its locus closer to the origin. This slows down the response of the closed loop system to meet the new settling time specification. Based on locus of the closed loop system, it is worth noting that this also reduces the gain reduced the system's overshoot and oscillations as the dominant poles are located on the real axis.

A. **Overshoot**

The percentage overshoot for the voltage loop was based on a single Li-ion cell within the load battery. The nominal charging voltage of a single Li-ion cell is $3.6V$, while the maximum voltage it can withstand – for a short time – without damage, is $4.2V$. Based on the following equation:

$$\%OS \leq \frac{V_{max} - V_{nom}}{V_{nom}} \quad (5.15)$$

Where V_{max} is the maximum charging voltage; V_{nom} the nominal charging voltage. Based on these values the overshoot specified was 17%. It is worth noting that this value is lower than its simulation equivalent. The major implication of this is that the phase margin requirement, resulting from the overshoot, is less. The current loop specification remained the same as the simulation results.

B. **Phase margin**

Practically, there are several factors that can introduce phase error in the voltage loop; for example, under heavy load conditions, the inductor can heat up significantly – this will change the properties of the inductor – this is because heating reduces the magnetic field strength and hence, inductance value. A smaller inductance value contributes to underdamping and increased overshoot and thus, decreased phase margin. Both input and output capacitors are also affected by temperature, however the AC power carried by both the switching EMI and impedance spectroscopy induced current response signals, is negligibly smaller, being only 4% of the DC power; since capacitors don't transfer DC power, the impact of changes in capacitance on the phase and, thus stability of the system is minimal. For the current loop, the transducers introduce negligible phase error in the system, as discussed. The filters introduce a phase error of 0.095° – a negligible delay. The phase margins were thus, chosen to be, 78° and 70° for the voltage and current loops, respectively – larger than the simulation margin because of the uncertainties in the phase delays than can be introduced by changes in the practical system, as discussed.

C. **Gain margin**

There are several factors that can contribute to gain uncertainties in a boost converter, and thus the uncertainty in the stability of the system. The most important factor is the converter's inductor; it stores its energy in its magnetic fields; under heavy loads its magnetic field strength decreases and thus, its energy storage capacity. If the inductor stores less energy, it is transferring less power i.e., lower voltage transfer ratio (lower gain). When assessing the practical system, it was observed that there was a, 4.5% average shift between the expected gain at the time-averaged duty ratio, and the resulting gain. Based on the voltage transfer ratio of a boost converter, this is an average gain uncertainty of $0.4dB$. In the current loop, the gain margin was chosen based on the

observed variability in the system at lower excitation frequency gain because of the solar module's varying parameters.

ii. *Steady state voltage results*

To achieve the required insolation and temperature levels for the design specifications, the solar simulating light board was utilized accordingly, as mentioned in the previous section. The module's input parameters were set at nominal conditions, unless stated otherwise. It is worth noting that the timescale is in microseconds, this was due to the syncing issues between MATLAB environment and the practical system, thus each large block in the grid represents 100ms and each smaller block represents 20ms.

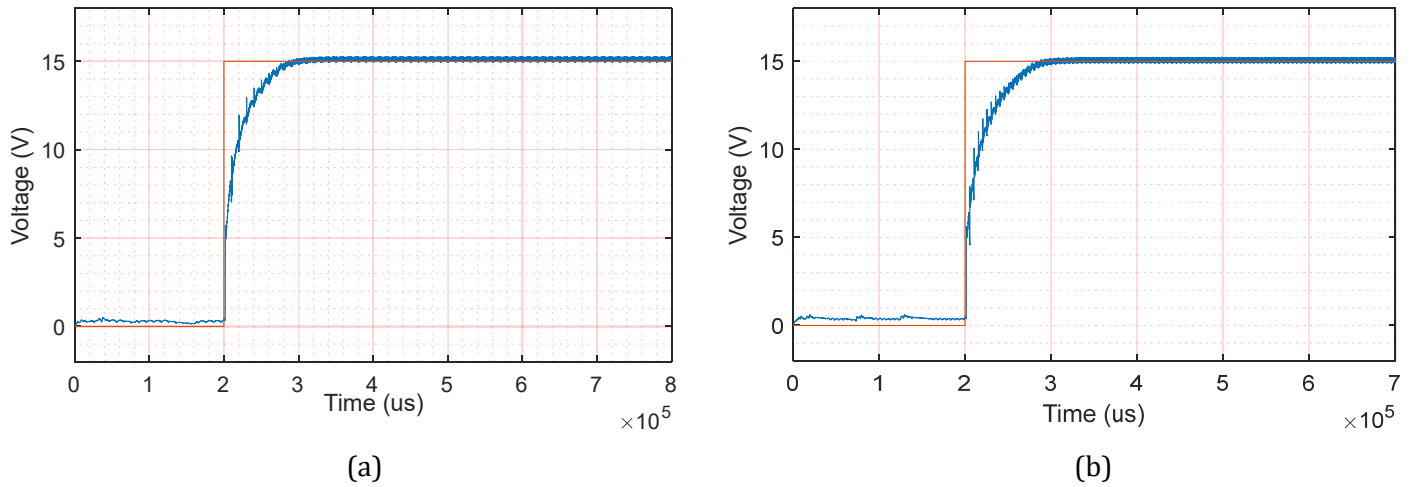


Figure 5.13 Steady state boost converter input and output voltage (a) single phase (b) multiphase

Figure 5.13 shows that both single-phase and multiphase responses have slowed considerably compared to their simulated counterparts, as designed for, now settling in 100ms – still meeting the practical specifications. The resulting waveforms now show overshoot with no oscillations. Figure 5.13 (a) shows a higher ripple factor than Figure 5.13 (b), as will be discussed in a later section. Both responses show good steady state errors of less than 1V.

iii. *Output voltage tracking specification*

The tracking specification was designed to assess the ability of the converter to respond to voltage setpoint commands in applications where the charging voltage needs to be adjusted in case of a failure demand change. The charging voltage command for the was stepped 15V to 10V, and the output voltage was observed. Figure 5.14 shows that the system exhibits good tracking for both commands. At lower voltages, however, the system has a poor transient response. This is largely due to the control action oscillating when attempting to operate at lower duty cycles; at these duty cycles, the switching utilization is poor. Table 5.2 shows that the performance of both

commands is within the desired specifications. Although the 10V has a larger output voltage ripple, it is still within a reasonable margin.

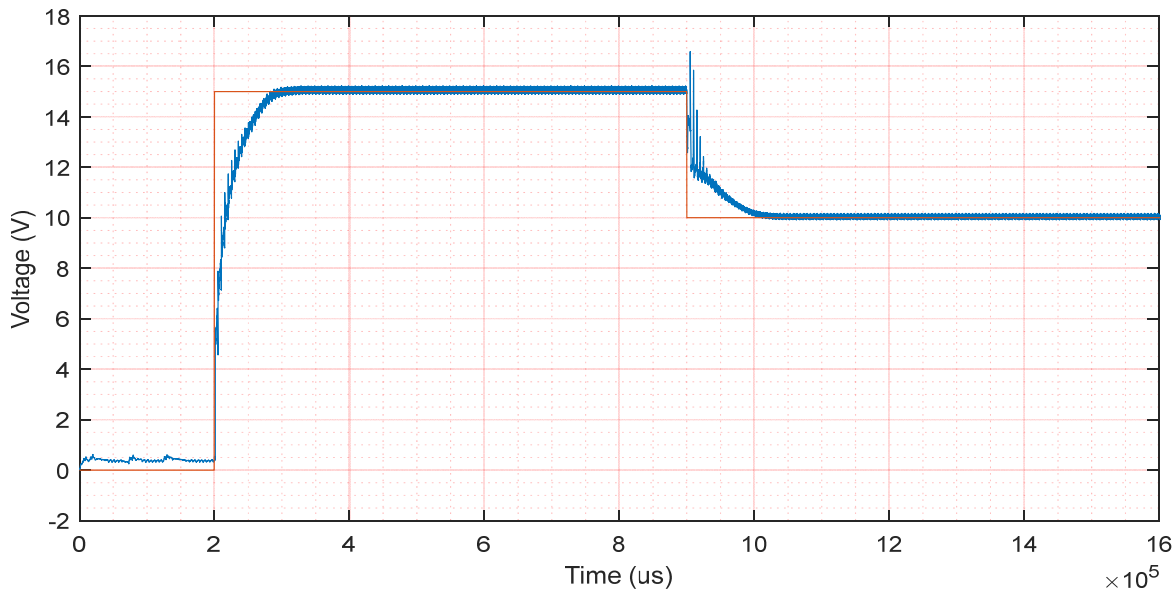


Figure 5.14 Output voltage performance under voltage setpoint commands

Table 5.2 Output voltage performance comparison under several voltage setpoint commands

Voltage command	Steady state error (V)	% Ripple
15V	0.21	4.6
10V	0.45	2.2

iv. **Load step**

The load current was stepped from the nominal charging current to the rated charging current of 25% of the battery’s capacity.

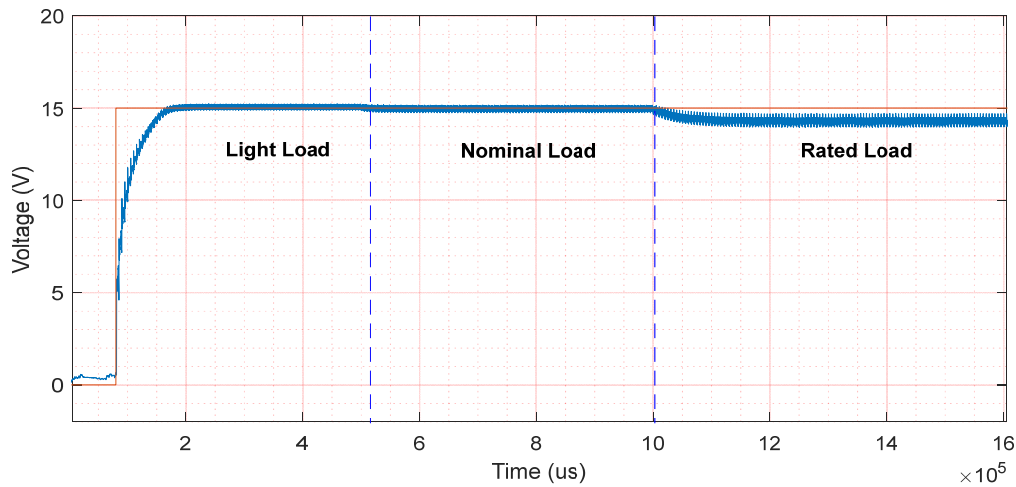


Figure 5.15 Output voltage performance under load disturbance

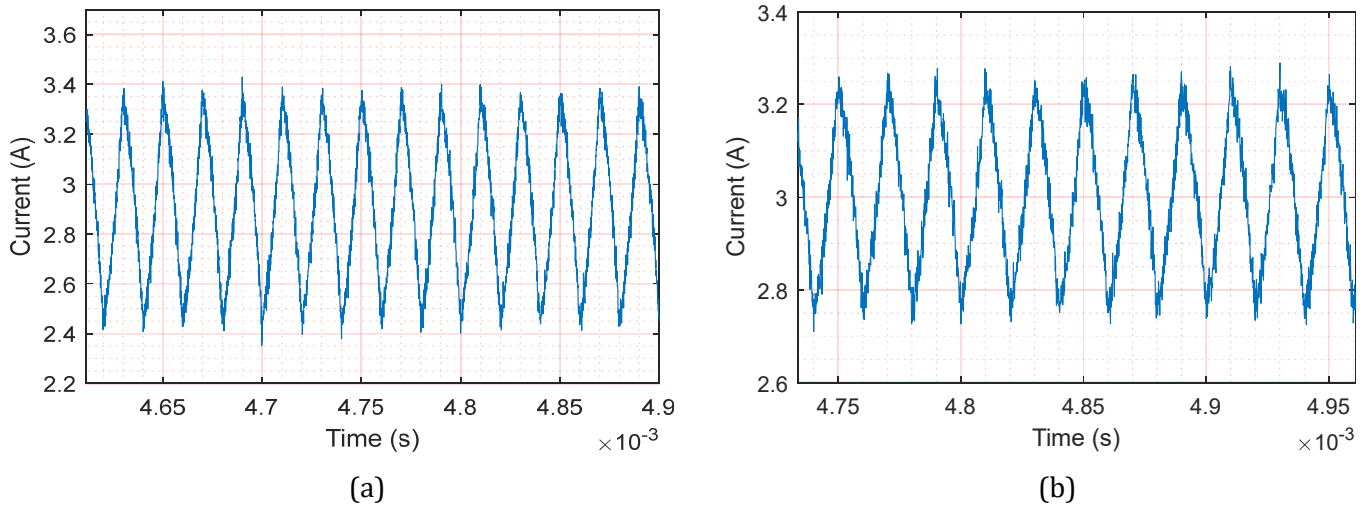
Table 5.3 Output voltage performance comparison under load disturbance test

Battery state	Steady state error (V)	% Ripple
Light load	0.21	1.93
Nominal load	-0.78	3.33
Rated load	-1.32	6.67

The battery was allowed to charge up such that its voltage was at the boundary between the pre-charge and constant current states, the load current was subsequently stepped up and charging voltage monitored. Figure 5.15 and Table 5.3 show good performance under light and nominal loads; at rated load, however, the voltage dips noticeably compared to lighter loads. The table shows that this results in the output voltage ripple exceeding the desired specification. This does not affect impedance measurements but, does impact the load power.

v. ***Input current analysis***

The peak-to-peak input current ripple for both converter configurations was assessed at the system's nominal load. Due to the availability of a single wide-bandwidth current transducer, only the input current ripple was measured for the multiphase, instead of the current from each of the inductors, as well the input current ripple, as performed during simulation results. The results are discussed in Figure 5.16.

**Figure 5.16 Boost converter input current ripple (a) single phase (b) multiphase**

The input voltage ripple from the single-phase converter is 33%, while the ripple from the multiphase is 13% of the module's output current. It is worth noting that both these results are within the desired specifications for the converter.

vi. **System efficiency analysis**

To perform optimal efficiency analysis on the system, the load battery was replaced with an electronic load; this was to ensure a constant load with each measurement. With the module output voltage kept constant at 17V, the load current was varied from a minimum charging current of 100mA, to the maximum charging current of 5.31A. This test was done to assess the performance of the converter at different charging currents and hence, load power, $P_o(W)$. Table 5.4 shows the efficiency characteristics of a boost converter as the load current varies, while Table 5.5 shows the performance of a multiphase converter. Table 5.5 shows that both single-phase and multiphase converters exhibit similar efficiency profiles. The single-phase converter does better at lighter loads specifically because the multiphase has higher inductive inertia, as well as higher losses in the form of voltage drops and parasitic component count. The multiphase converter does, however, offer better performance at heavier loads than the single-phase converter, with peak efficiencies above 70% compared to efficiencies in the high 60s for the single-phase converter. These efficiencies are, however, lower than the efficiencies observed in similar converter-based research [1] – [3]. This can be mainly attributed factors such as, high switching frequencies, losses along the DC cables, and the skin effect. In similar research, switching frequencies not larger than 50kHz were employed, as opposed to 500kHz for this research. Both converters do struggle with maintaining high efficiencies under rated load conditions, no doubt from the effects discussed.

Table 5.4 Efficiency analysis of a single-phase boost converter

$I_{pv}(A)$	$V_{pv}(V)$	$P_{pv}(W)$	$I_o(A)$	$V_o(V)$	$P_o(W)$	% <i>eff</i>
0.59	17.01	10.04	0.30	18.71	5.52	0.55
1.12	16.97	19.01	0.56	20.36	11.40	0.60
1.49	17.02	25.36	0.75	22.13	16.48	0.65
2.21	17.00	37.57	1.11	23.80	26.30	0.70
2.99	17.05	50.98	1.50	24.21	36.20	0.71
3.61	16.99	61.33	1.81	22.26	40.17	0.66
3.89	16.95	65.94	1.95	21.59	42.22	0.64
4.11	17.09	70.24	2.06	21.41	44.10	0.63
4.51	17.05	76.90	2.26	20.40	45.03	0.60
5.12	17.03	79.60	2.56	18.36	46.17	0.55

Table 5.5 Efficiency analysis of a multiphase boost converter

$I_{pv}(A)$	$V_{pv}(V)$	$P_{pv}(W)$	$I_o(A)$	$V_o(V)$	$P_o(W)$	% <i>eff</i>
0.53	17.03	9.03	0.27	16.69	4.42	0.49
1.01	17.11	17.28	0.51	18.82	9.50	0.55
1.55	17.06	26.44	0.78	22.18	17.19	0.65
2.05	16.99	34.83	1.03	23.79	24.38	0.70
2.61	16.93	44.19	1.31	25.06	32.70	0.74
3.23	17.01	54.94	1.62	24.77	40.02	0.73
3.55	17.02	60.42	1.78	23.66	42.11	0.70
4.11	17.06	70.12	2.06	21.90	44.36	0.64
4.514	16.98	76.65	2.26	20.82	45.12	0.61
5.12	17	79.84	2.56	18.75	46.31	0.54

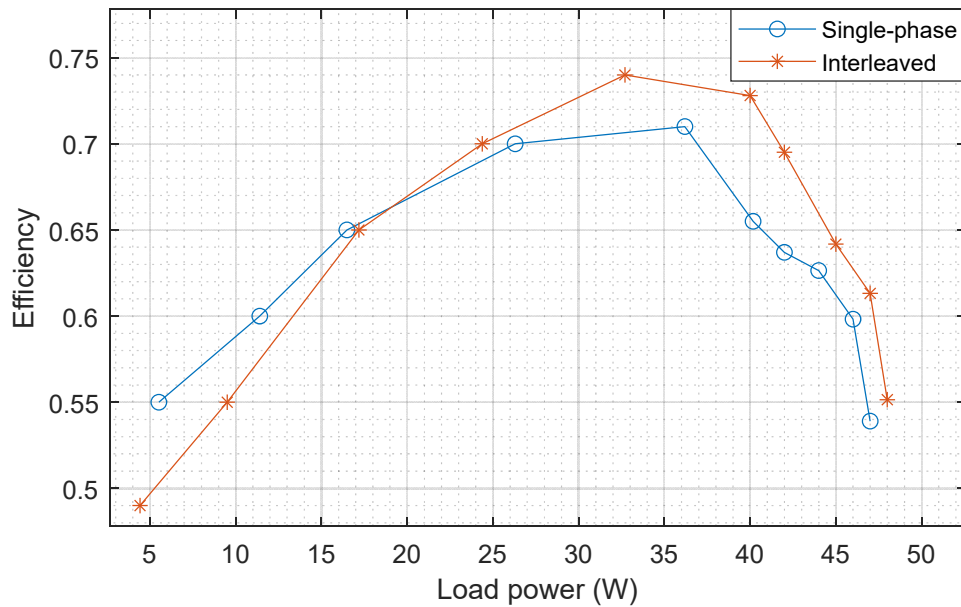
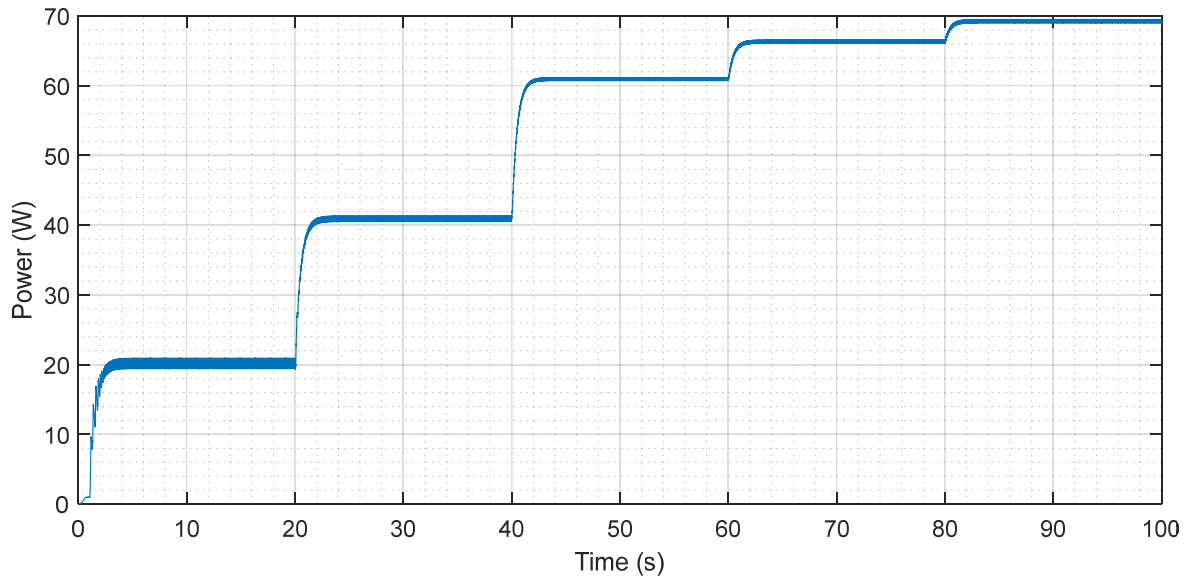


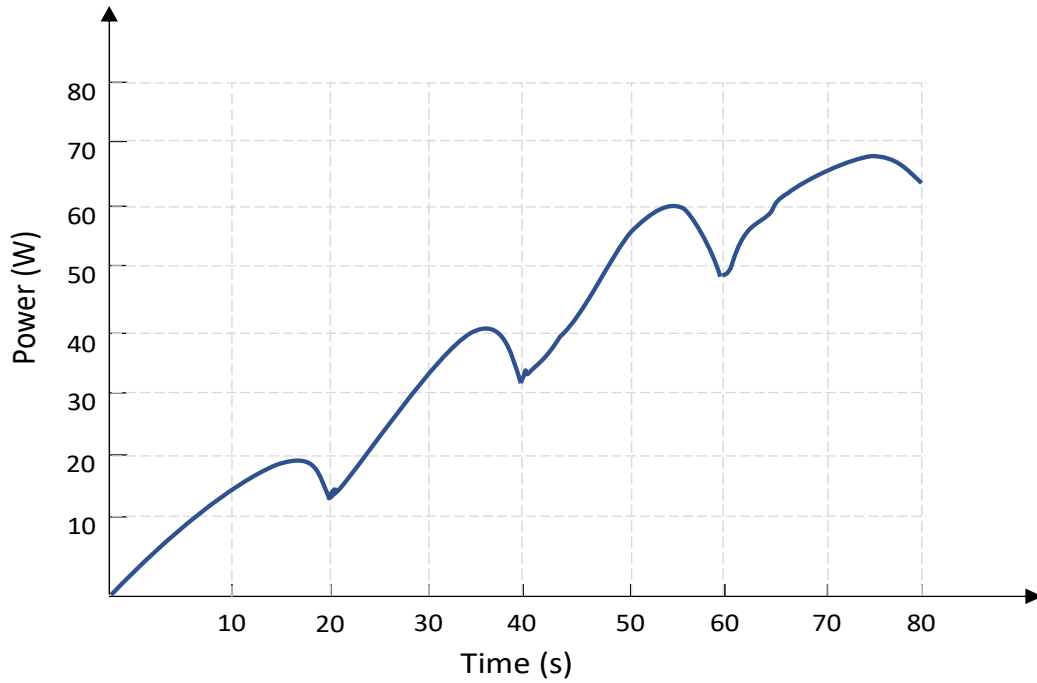
Figure 5.17 Comparative efficiency analysis between single-phase and multiphase converter with varying load power

5.6.4 Maximum power point tracking

When performing maximum power point tracking through a switch-mode converter, the module treats the converter as a variable electronic, where assuming small voltage steps, the converter can be considered a linear load. Various irradiance values were imposed upon the module; Figure 5.18 shows the subsequent maximum at each irradiance level. The NI DAQ and MATLAB were used to track the maximum power point at 20s intervals.



(a)



(b)

Figure 5.18 Module incremental conductance power point tracking under varying irradiance (a) steady state (b) instantaneous

Figure 5.18 shows that the algorithm can track maximum power points over various irradiance levels. There are oscillations at the maximum power point, not exceeding $3W$ at each maximum power point. The settling time for algorithm is, $\approx 2s$; that is a quick convergence time adequate for rapidly varying conditions. Evidently, the lower the maximum power the larger the oscillations. Although the algorithm exhibits such oscillations, they are

significantly small compared to the average power at each level, as well as significantly less than the oscillations from a typical P&O algorithm.

5.6.5 *Module optimal excitation signal design and bandwidth characterization*

To ensure that module's impedance is estimated within the linear range of the module, the optimal amplitude was designed using the frequency response analyser. Configured in potentiostatic mode, various small voltage AC perturbation single frequency sinusoids were added to the module's operating point and the current response of the module monitored.

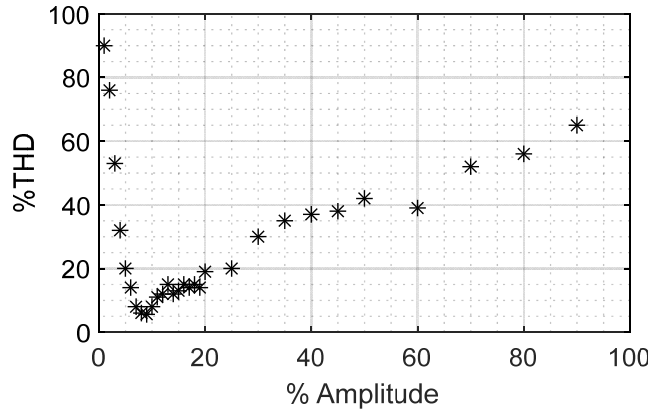


Figure 5.19 Response current THD resulting from voltage perturbations of amplitudes, as a percentage of the bias voltage

The THD of the response signals were then observed, as illustrated in Figure 5.19; this is because the THD indicates the system's non-linear behaviour. An amplitude range that was between, 8.51% and, 13.45% of the modules operating voltage, yielded the lowest THD. Voltage ranges below this range result in significantly low SNR which makes signal detection difficult at these ranges, thus increasing the THD as system noise is picked up by the analyser. Voltage ranges larger than the linear range result in a nonlinear system as these amplitudes move the system toward the module's voltage source region. The above-mentioned linear range was then chosen to ensure that the system remained linear during the measurements.

The bandwidth of the module was subsequently determined by performing EIS on the module and observing the Nyquist plot of the module's impedance. The observations noted the higher and lower frequencies where the device started to conduct i.e., frequencies where the imaginary impedance components are positive. The optimal bandwidth for module was found to be in the range, $[9.71; 109]kHz$ – the range rounded down to $[10; 100]kHz$ because there was no significant change in the impedance between rounded range and the actual range.

5.6.6 *Impedance spectroscopy experimental voltage excitation signal induction*

From the simulation results, 3 excitation signals (chirp, PRBS and triangular wave) were chosen for purposes of device excitation, as discussed in the previous chapter. These signals were thus assessed based on the

specifications. The excitation signals were also assessed at multiple operating points to analyse characteristics of the signal.

i. Experimental considerations

During the simulation results, AC EEC parameters based on a single operating point for the module were chosen, as mentioned. In the experimental results however, the actual module was used. The conditions were controlled such that a single operating point (matching the AC EEC parameters used in simulation) could be maintained as accurately as possible. The biggest factor that impacts the accuracy of the single operating point is the temperature, since the solar light board was maintained at the same illumination during the results. Thus, between each measurement the module was allowed to cool before the next measurement. Although this significantly impacts the data acquisition time, it helps maintain a high accuracy in the measurements.

The ADC of the NI DAQ is 16bits, as discussed, hence, the maximum SNR for each of the perturbation signals is: $-20\log_{10}\left(\frac{1}{2}\right)^k$; where, k is the resolution of the ADC in bits; for a 16bit ADC this is a maximum SNR of, 96.32dB. The perturbation duty cycle signals have finite resolution under experimental excitation. Thus, the maximum number of clock cycle available per period is, $\delta = f_{ck} \cdot (f_{sw})^{-1}$; where f_{ck} is the DAQ's maximum clock frequency of 20MHz, and f_{sw} is the converter's maximum switching frequency, of 500kHz. This means that the carrier can be implemented, " $\delta/2$ " times if a triangular wave carrier signal is used. Hence, the duty cycle resolution $\partial d_{in} = \delta^{-1}$, or 5×10^{-3} . Consequently, there is a maximum of 2.6 possible voltage steps to approximate the maximum perturbation duty cycle of 0.13. This is clearly not sufficient for digital to analog approximation. To improve the resolution in the perturbation signal, a sawtooth carrier was chosen instead, this ensures that the carrier and hence, perturbation signal has, " δ " discrete values, i.e., the duty cycle resolution proves to 25×10^{-3} . This results into a maximum of 5.2 possible voltage steps to approximate the maximum perturbation duty cycle. Although this is an improvement, 5.2 levels are still considerably low for signal reconstruction. The use of multiphase converter can help improve this resolution by averaging the duty ratios through each of the phases. Because the inductances used in the design of the 2-phase multiphase converter are the same, the analysis of the effective averaged duty ratio becomes, $d_{eff} = d + \frac{\partial d_{in}}{2} + \frac{r_{La}I_{dc}}{2V_{dc}}$. This means that with a 2-phase multiphase converter, the duty cycle resolution can be improved to 12.5×10^{-3} , resulting in 10.4 levels – considered sufficient in signal approximation. As mentioned earlier in the chapter, the NI DAQ has a maximum of $2MS.s^{-1}.ch^{-1}$; this means that the product of the sampling frequency chosen, as well as the number of samples must, at any given time, be equal to 2MS. Therefore, the trade-off between the sampling frequency and the number samples will be chosen on a per-signal basis, to achieve the most optimal signal resolution. The other factor to consider when choosing the ADC's sampling rate is the aliasing effect. When acquiring either perturbation, excitation or response signals, discrete Fourier transform (DFT) is performed on the signals for their magnitude and phase

characteristics. The transform will yield the required output provided that there is no aliasing effect. That is, the ADC's sampling frequency must be at least twice the maximum impedance frequency – in the case of the module used in this study, 100kHz – this is known as the Nyquist-Shannon sampling theorem. Hence, as observed from the system setup, low pass filters were implemented to ensure that high frequency noise was eliminated from the signals to avoid aliasing.

Another consideration to consider is the excitation amplitude of the induced voltage excitation signal. From the excitation specifications, a module operating at an operating voltage of that is 78% of the module's $17V_{mp}$, will need a minimum excitation amplitude of $1.13V$ and a maximum of $1.78V$. The lowest impedance magnitude is observed at lower frequencies, as will be discussed during the results section. At a frequency of 10Hz , the module has an impedance magnitude of $630m\Omega$; this means that at this perturbation frequency, the current induced at the module by the excitation voltage would be, $2.83A$; this amount of current would not only move the system into a nonlinear operating range but would also increase the operating temperature of the system. Hence, as discussed in section 5.4.5, the module's operating voltage was reduced to 40% of the module's maximum voltage, this reduced the maximum induced current to $800mA$, sufficient at keeping the system within its linear range.

ii. *Chirp excitation*

When generating a chirp perturbation signal, the number of samples take priority. With 10 samples chosen, a maximum sampling frequency of 200kHz was possible (meeting the minimum requirements to avoid aliasing); this clearly shows that with the NI DAQ, producing an analog signal at 100kHz is challenging. This is, however, a limitation of the available processing equipment, and the implications of this will be assessed in the impedance measurements taken with the signal, later in Chapter 7. The results for the voltage excitation and current response signal are analysed below.

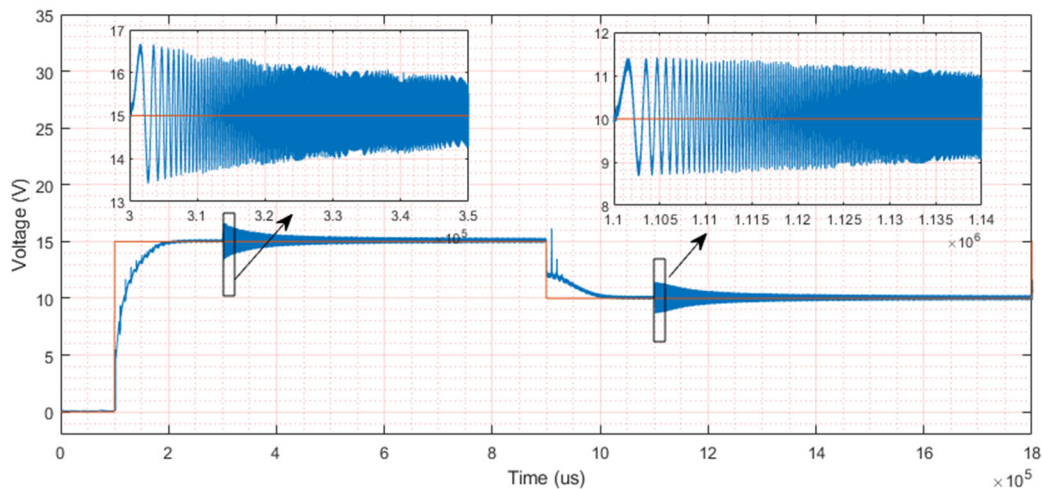


Figure 5.20 Induced chirp voltage excitation signal under input voltage variation

Figure 5.20 shows that the induced voltage excitation signal in, both voltage commands, has large amplitudes at lower frequencies and small amplitudes at high frequencies, as hypothesised. At 15V, the lowest amplitude is 20% of the module's output voltage, while the highest frequency is 1.3% - both of which lie outside the desired range. This leads to a crest factor of 3.4 and an effective excitation bandwidth of $[3.5, 24.6]kHz$. The 10V command, the lowest frequency is 28.5%, while the highest frequency is 3.75% - with a crest factor of 4.8 and an effective excitation frequency of $[1.7, 18.6]kHz$. The SNRs are 67dB and 61dB for the 15V and 10V commands, respectively. Figure 5.21 shows that the induced excitation signals, for all 3 load conditions are as discussed earlier. The crest factor is still significantly large, with a reduced excitation frequency. Taking measurements under rated load conditions, is not ideal because the system oscillates significantly.

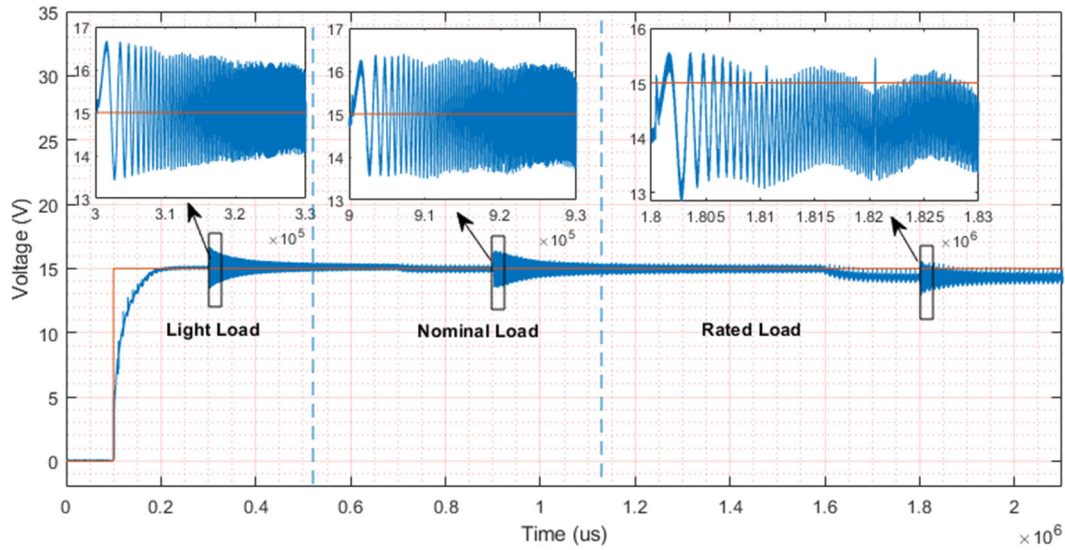


Figure 5.21 Induced chirp voltage excitation signal under load variation

iii. *PRBS excitation*

The sequence, being a digital signal, means that a few samples is required for signal generation; this means high frequency resolution was attainable. This is an important factor, as the simulation results showed that high frequency impedance estimation proved challenging through PRBS induced excitation. With 8 samples chosen, a maximum sampling frequency of 250kHz was possible. This does however, mean that high frequency resolution is poorer than desired. With 5 samples chosen, a frequency of 400kHz achieves a better resolution at the expense of a larger quantization error. Based on the quantization error analysis, the quantization error difference between 5 and 8 samples, is 4%, an error considered within an allowable margin of error. The results of the perturbation, induced voltage excitation, and current response signals are discussed below.

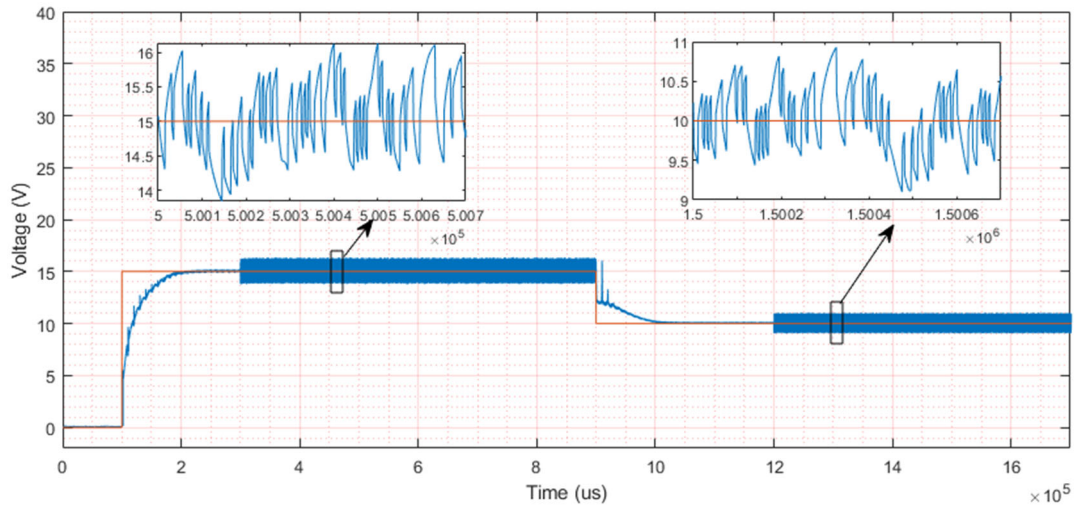


Figure 5.22 Induced PRBS voltage excitation signal under input voltage variation

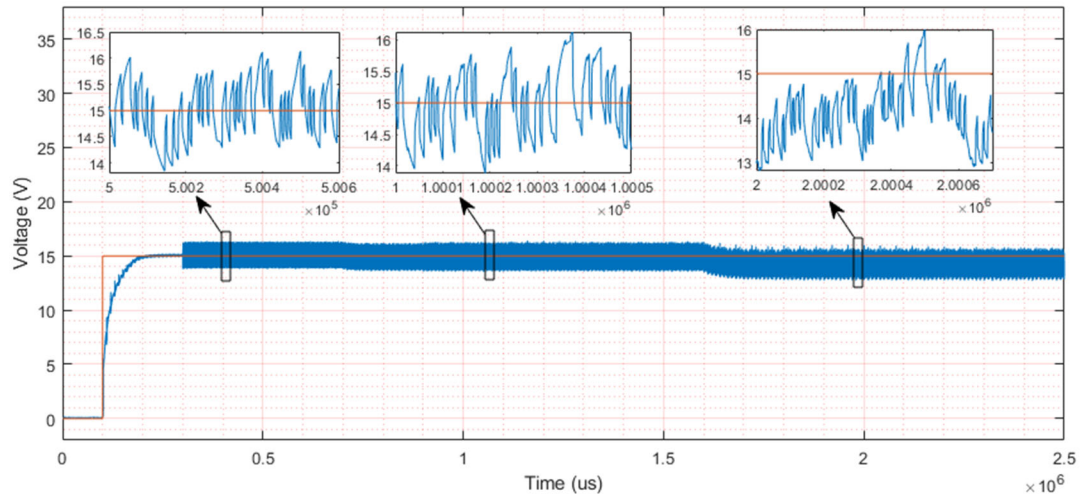


Figure 5.23 Induced PRBS voltage excitation signal under load variation

Figure 5.22 shows that, for both voltage commands, the induced voltage excitation signal due to PRBS perturbation signals lie outside the designed amplitude range. They both show expected lower amplitudes at higher frequencies and higher amplitudes at lower frequencies. The 15V step has a 4.1 crest factor, while the 10V step has a 4.3 crest factor – both outside the desired specifications. The SNRs are 64dB and 59dB for the 15V and 10V commands, respectively. Figure 5.23, on the other hand, shows similar performance to the chirp excitation because at rated load, the voltage exhibits the highest crest factor and excitation amplitude, leading to nonlinear behaviour in the module.

iv. ***Triangular wave excitation***

The triangular voltage excitation signal does not suffer from issues of quantization in its generation because it is generated by the converter. The converter is an analog device and thus has infinite resolution. The limitation to

the signal is introduced when processing both voltage and current signals for impedance estimation because the ADC is resolution limited, as discussed earlier in the section. As discussed during the simulation results, the excitation frequency of the inductor current ripple was limited to, 20kHz , the lowest optimal switching frequency for the converter; the implications of this on solar module impedance measurements will be evaluated in Chapter 9. It is worth noting a single-phase boost converter was used for these results. The inductor current ripple and the module's voltage response are discussed. Figure 5.24 shows that the current ripple for the 20kHz switching speed is 49% of the DC current, while it is 33% for the 100kHz switching speed. This large ripple values show that the design of converter in its current state is not adequate for excitation or power regulation. In Chapter 9, a through approach to this type will be undertaken to ensure that the excitation signals meet the designed specifications.

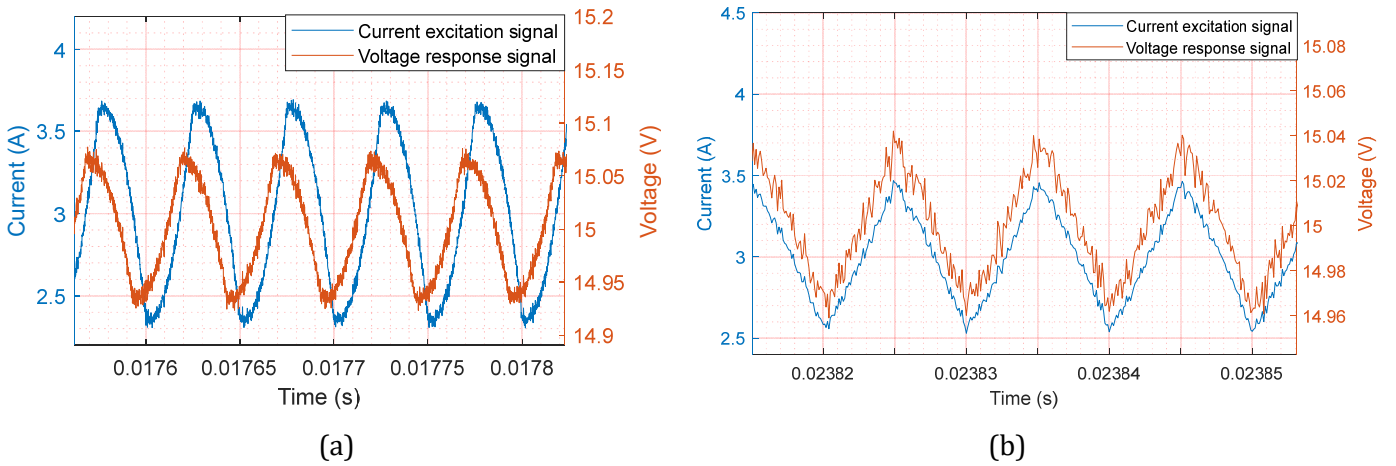


Figure 5.24 Solar module inductor current excitation (a) 20kHz (b) 100kHz

5.6.7 Offline Solar module impedance estimation results performed using a frequency response analyser

The FRA was used to characterize a solar module under several operating condition; this was done to create a results benchmark with which the performance of the designed online impedance extraction system will be assessed.

i. Varying operating point under constant irradiance and temperature

The first characterization was taken under various operating points while the module had the same irradiance and temperature values. String inverters have a command function to curtail the power generated by the modules depending on the load demands. If the modules are producing more power than required, the inverter sends a command to reduce the PV module's power output; this is achieved by changing the voltage bias, and hence operating point of the solar module strings – this process is referred to as power curtailment. The results presented in this section thus, emulate and analyse the impedance of a single solar module under various power

curtailment commands of a solar power plant. The FRA has a command that allows for voltage bias on the module to varied; this change is better illustrated by Figure 5.25. The figure shows operating points in colours that are matched with the impedance curves for better illustration.

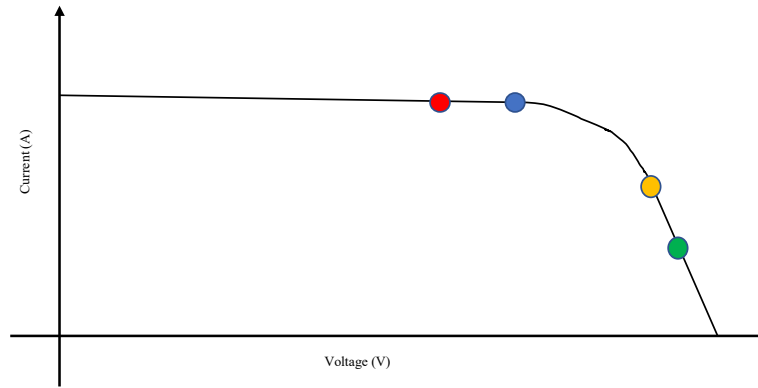


Figure 5.25 Points on the I/V curve at which varying operating point impedance measurements were taken

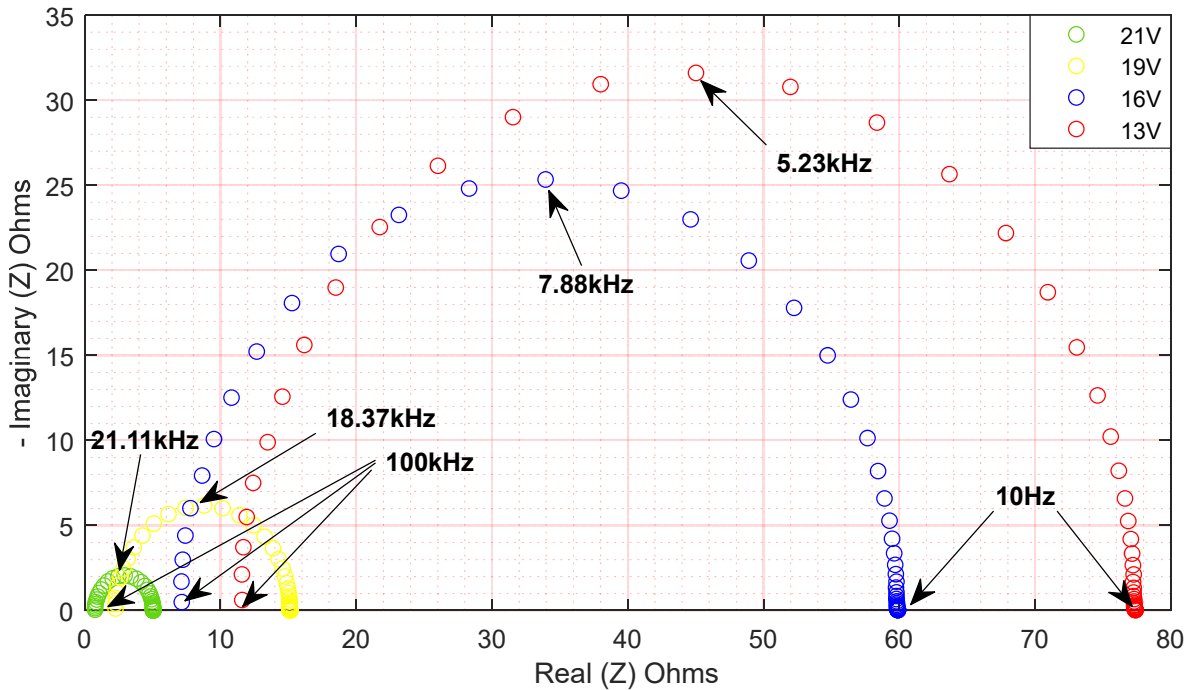


Figure 5.26 Module Nyquist impedance response under high irradiance conditions and varying temperature

Table 5.6 Monocrystalline solar module parameter estimation under variable operating points

Parameter	13V	16V	19V	21V
r_s	12.3Ω	7.98Ω	2.69Ω	210mΩ
r_{np}	64.5Ω	52.1Ω	13.5Ω	5.12Ω
c_{np}	4.58mF	6.24mF	24.4mF	59.0mF
r_{pp+}	—	—	—	—
c_{pp+}	—	—	—	—

Figure 5.26 shows that, provided the operational conditions remain constant, a higher operating voltage decreases the module's impedance magnitude profile. This, however, does not apply to individual parameters as illustrated in Table 5.6. The module's series and diffusion resistances decrease with increasing operating voltage; this can be because with increasing junction forward bias voltage, the resistance in the junction must decrease to maintain the volume of charge flowing through the junction and load. The junction capacitance, does, however increase with the bias voltage; in forward bias mode, this capacitance is much larger than the module's junction capacitance. As discussed earlier, the diffusion capacitance occurs due to charge carriers stored around the depletion region; thus, when a forward bias voltage is applied, the majority carriers in the N region migrate to the P region and recombine with holes, while the minority charge carrier migrate to the N region. When this happens, the width of the depletion zone decreases; but capacitance is inversely proportional to its width, thus explains the increase in the diffusion capacitance.

ii. Single operating point under varying irradiance and constant temperature

For the purposes of this study, irradiance levels, $> 600 \text{ W/m}^2$ are considered well-lit, while irradiance levels between 100 W/m^2 and 600 W/m^2 are considered low-lit – any lighting conditions below 100 W/m^2 are considered dark conditions in the context of using a power converter. During each of the curves the module's operating temperature was allowed to reach a terminal value of 31°C , before the measurements were taken. The terminal value of the chosen temperature was high because to achieve high irradiances, the distance between the light board and module should be small. Two irradiance levels were tested under well-lit conditions, while 2 irradiance levels were taken under low irradiance conditions. Figure 5.27 shows the illustration of the points at which the impedances were taken; the colours were also matched to the Nyquist plot. The measurements were taken in module's current source region because with irradiance variations, that region experiences significant changes.

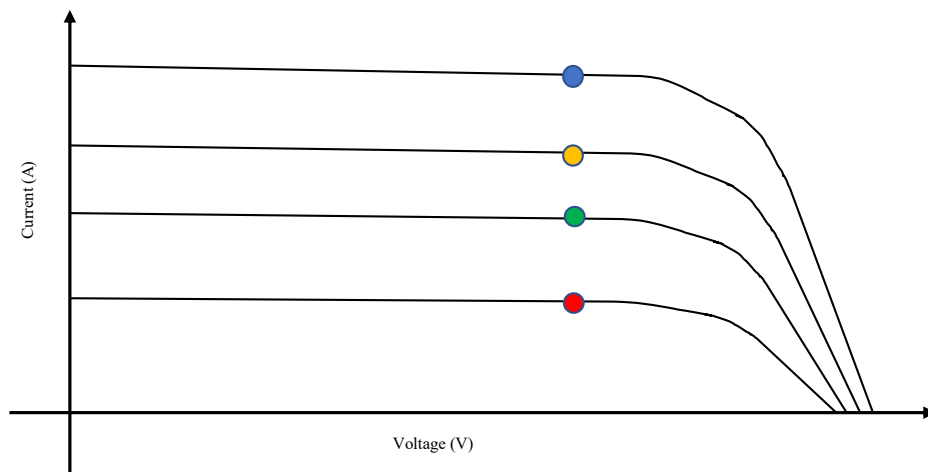


Figure 5.27 Points on the I/V curve at which varying irradiance impedance measurements were taken

Figure 5.28 shows that under good lighting, the series resistance of the module largest, at $215m\Omega$ and $133m\Omega$ for the $730 W/m^2$ and $602 W/m^2$ irradiances, respectively; this results from the larger number of electrons flowing through the cells and contacts under high irradiances. The diffusion parameters decrease with increasing irradiance, however, as a result of the decreasing light energy and hence, decreasing rate of electron generation. It is worth nothing that these results are in line with results obtained in [4], which investigated multiple cases of operational and fault conditions; as well as [5] and [6]. The FRA chooses its own frequencies within the specified min and max values. For high irradiance conditions, there are several frequencies at high frequencies that return an error, resulting from a lack of impedance convergence by the software. This is largely due to non-linear effects that the FRA picks up from electron interference, and then rejects the frequencies. The research did not, however, evaluate the excitation frequency distribution and its significance on the module's characterization. The assessment of the results shows that the frequency distribution of the impedance on a Nyquist plot is largely dependent on the irradiance levels. Evidently, under these conditions, the module responds better to higher frequency perturbations; that is, frequencies $> 20kHz$ make up at least half of the module's characteristics, while the frequencies $> 1kHz$, dominate the response. This information is useful is the selection and design of excitation signal characteristics, as will be discussed in later chapters. The $100kHz$, however, consistently appears at the Real axis intercept for the response; this shows that it is possible to fully characterize the module's series resistance just by assessing the module's response at that frequency. The majority carrier parameters, which appear at higher frequencies, were null; this can be due to the poor resolution at higher frequencies.

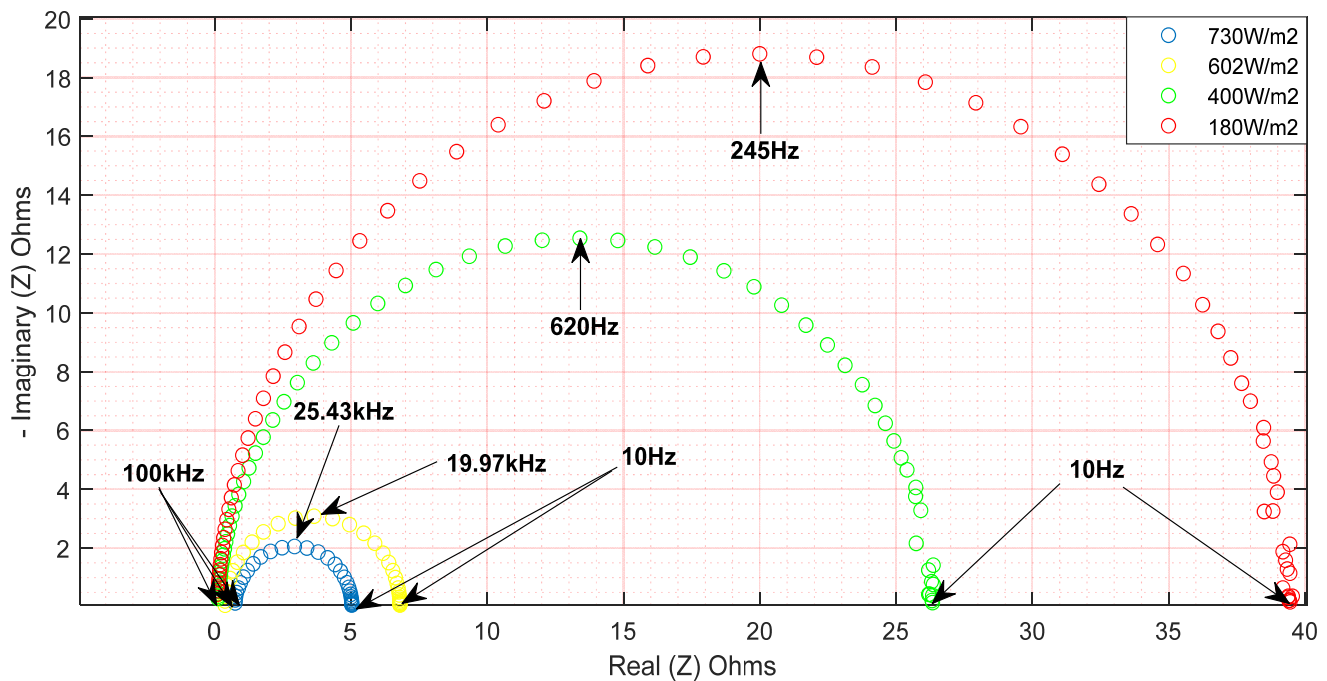


Figure 5.28 Module Nyquist impedance response under varying irradiance levels

Table 5.7 Monocrystalline solar module parameter estimation under varying irradiance

Parameter	180 W/m ²	400 W/m ²	602 W/m ²	730 W/m ²
r_s	98mΩ	125mΩ	979mΩ	1.56Ω
r_{np}	38.5Ω	22.1Ω	6.21Ω	3.75Ω
c_{np}	8.38mF	12.2mF	53.1mF	79.6mF
r_{pp+}	—	—	—	—
c_{pp+}	—	—	—	—

Figure 5.28 shows that, similar to high irradiance conditions, the module's impedance profile increases with decreasing irradiance, with the series resistance increasing with increasing irradiance. The curves do, however, show a better high frequency resolution compared to high irradiance conditions. The simplest explanation to this would be that there is too much electron-movement interference under high irradiance conditions which impacts higher frequency linearity – this explains the use of dark I/V curves. The plots show that, at 180 W/m², the peak response of the module occurs at 201Hz – 124 times lower compared to the 25.16kHz peak response from the module under high irradiance conditions. The relationship between the irradiance and frequency at the peak response, is therefore nonlinear. Although the 180 W/m² irradiance level has the highest impedance, it does exhibit the quickest junction time constant. On a solar cell level, this means that when light of low intensity levels strikes the depletion layer, the rate at which the electrons accumulate at the N-side of the junction, to develop at potential difference, is higher.

iii. Single operating point under varying temperature and constant irradiance

To understand the effects that temperature has on the module's impedance parameters, this assessment was undertaken. It is desirable to keep the irradiance levels constant during the measurements to ensure that the changes in the impedance measurements are only from temperature changes; however, changing the temperature on the module, by varying the distance between the light board and module, changes the irradiance as well. Hence, a blower fan was used to regulate the temperature. Figure 5.29 shows the operating points where the module's impedance was taken during temperature changes. It is worth noting that this is not the most effective temperature-control technique; it has a long settling time i.e., the time it takes the temperature to reach its final value when using the blower fan; the blower fan used, only had 3 speed control settings which limits the level of available control; and lastly, it presents a large uncertainty factor, not only in the actual module temperature, but the measured impedance as well. As such, the results presented for temperature represent the general temperature trend but, are by no means the most accurate representation of the effects that temperature has on the module's parameters at a single operating point.

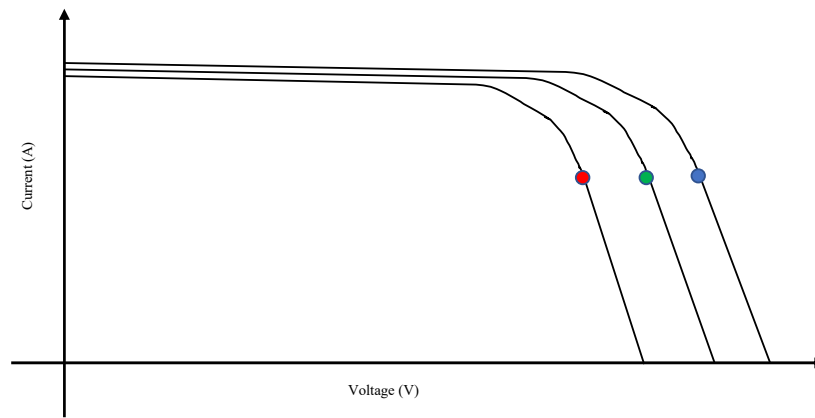


Figure 5.29 Points on the I/V curve at which varying temperature impedance measurements were taken

The measurements were taken in the voltage source region because as evident from the Figure 5.30, changes in the current source region are not as significant as changes in the voltage source region; this will ensure optimal impedance characterization due to temperature changes. Unlike the irradiance, Figure 5.30 shows that increasing temperatures degrades the module's performance; the figure reiterates that at high temperatures, the module's series resistance increases, while the diffusion capacitance decreases. The capacitance decreases because of the effect temperature has on capacitive parameters.

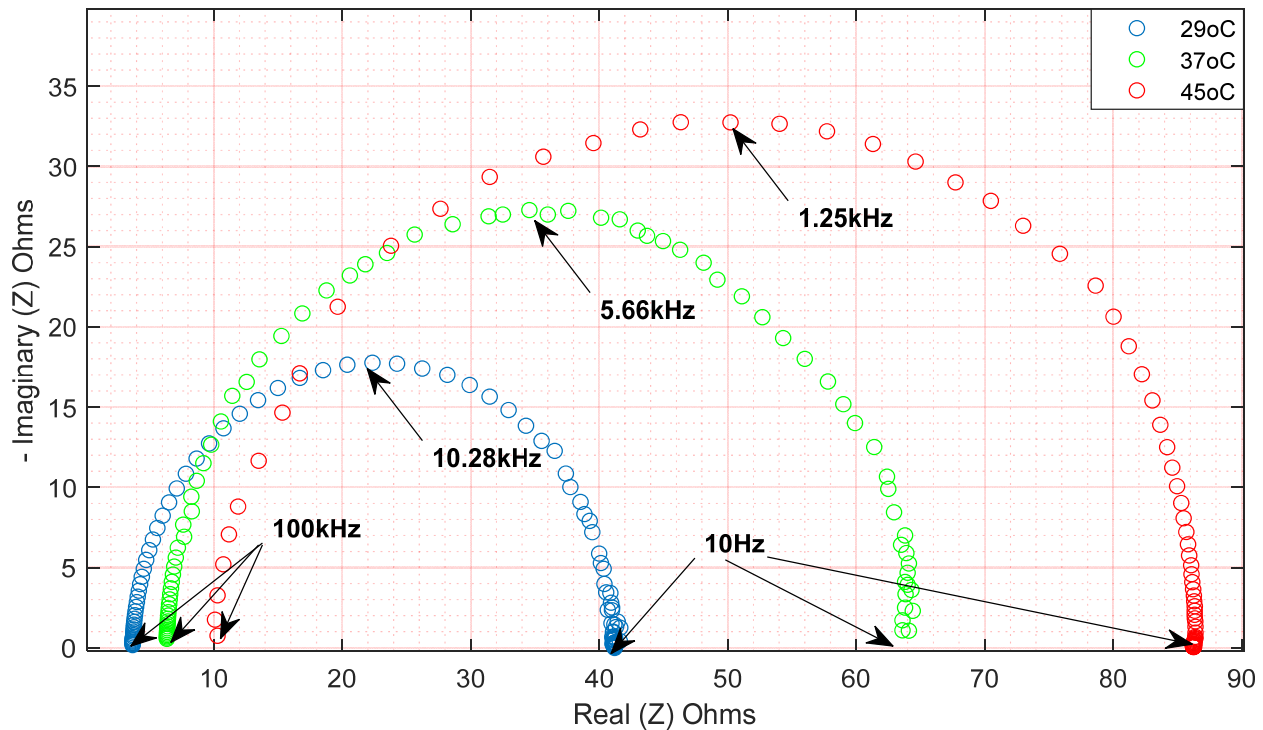


Figure 5.30 Nyquist plot showing temperature effects on the impedance of a solar module

Table 5.8 Monocrystalline solar module parameter estimation under temperature changes

Parameter	29°C	37°C	45°C
r_s	4.57Ω	5.98Ω	10.1Ω
r_{np}	38.7Ω	59.4Ω	76.89Ω
c_{np}	9.95mF	5.89mF	4.82mF
r_{pp+}	—	—	—
c_{pp+}	—	—	—

iv. Partial shading and hot spots under constant irradiance

The module was partially shaded at an irradiance level of 600 W/m^2 to emulate typical conditions in the field. The 36-cell module, used in this study, has 2 body diodes each bypassing an 18-cell series combination; therefore, there is only a single partial shading case activating the diode that can be investigated. This is because if both bypassing diodes are activated, the module is not partially shaded, but considered fully shaded. Under partial shading, the shaded string of cells is completely bypassed, thus, the impedance response is that of a series configuration between the body diode and the unshaded string of 18 cells. Figure 5.31 shows operating points from which the impedance values were taken.

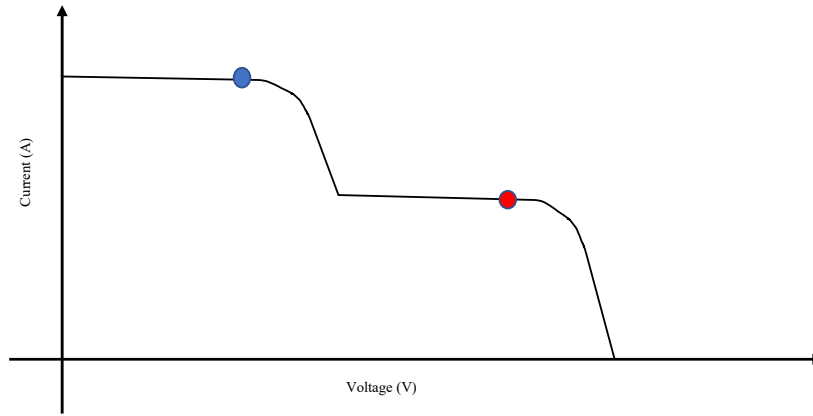


Figure 5.31 Points on the I/V curve at which partial shading impedance measurements were taken

As observed from Figure 5.32, partial shading impacts the module's operating current at high voltage biases, more specifically, reduced the maximum current by half, as expected. This is because the partially shaded cells dictate the module's maximum operating current. Under the partial shading, the response resembles that of a lower-lit module. The increase in the impedance is dependent upon the type of bypass diode used in the module. This means that the impedance response of the bypass diode used in this module is much larger than the module's cells, as it dominates the response.

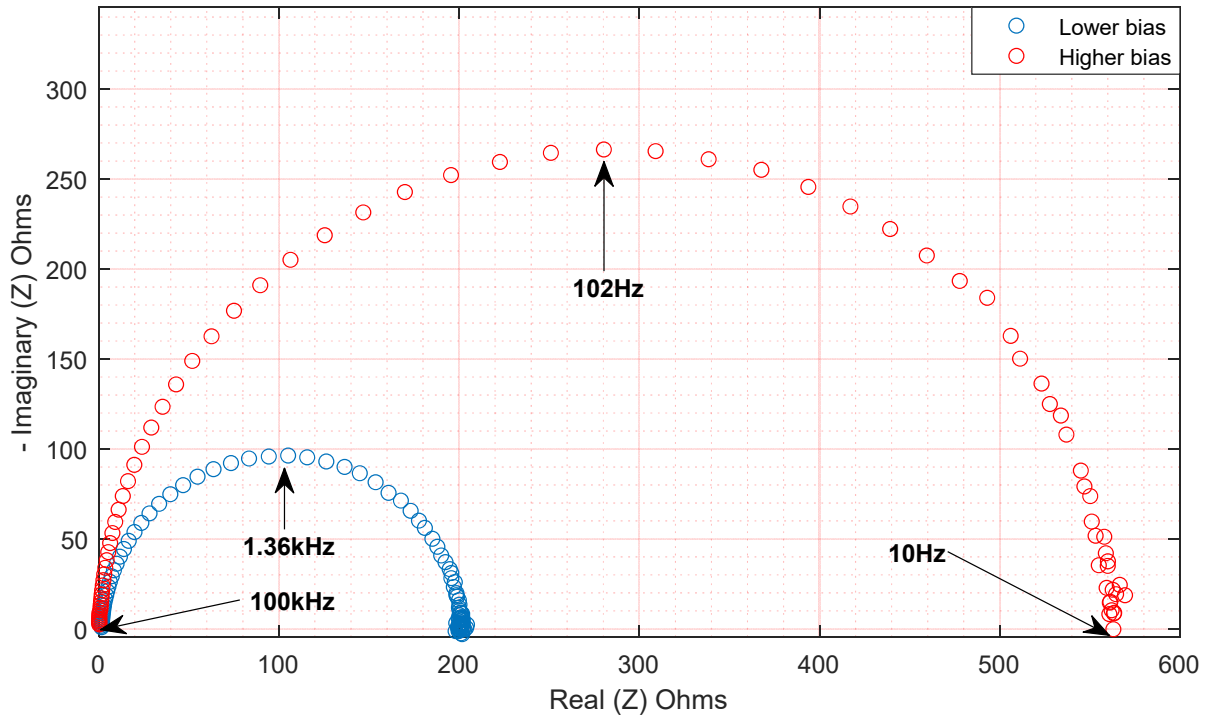


Figure 5.32 Module Nyquist impedance response under partial shading conditions

One thing to note is that although the overall response is dominated by the diode, the series resistance remains fairly low, consistent with low irradiance conditions: in fact, it is quoted as $523m\Omega$. At this irradiance the series resistance is expected to be, $\approx 133m\Omega$, thus the diode's series resistance is $390m\Omega$. The results are in line with the investigation undertaken in [7], where the research showed an increase in the impedance at high irradiances with partial shading cases. Hotspots are a consequence of partial shading; a partially shaded cell acts as a load and when the irradiance levels increase, the power from unshaded cells is dissipated in the shaded cell causing its internal temperature to rise. Hence, the observed characteristics that resemble temperature variation.

v. *Micro cracks*

The microcracks were performed on a $45W$ polycrystalline solar module, as mentioned in the introduction. The module's micro-cracks were introduced using a lab iron. The module's maximum output voltage was larger the converter's required minimum input voltage thus, making the module viable for characterization under the designed system. The assessment was undertaken under well-lit conditions; the impedance is illustrated in Figure 5.33. Results show that microcracks have very little effect on the module's overall impedance, except for series resistance – a similar effect to partial shading and hotspots; that is, under well-lit conditions, the impedance is expected to be low, however, as observed from the results, the impedance shows a slight increase. For microcracks this can be explained by considering the assembly of a single solar cell; during a crack, the metal

finger contacts used to maximise current collection are fractured or broken; when this happens amount of current collected by the cell decreases.

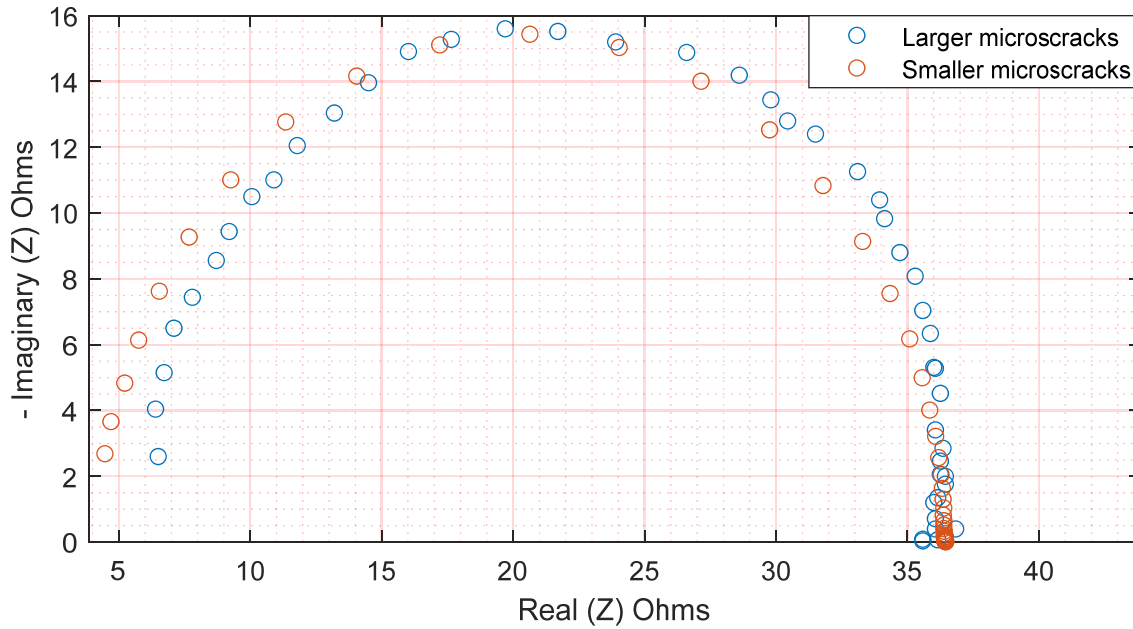


Figure 5.33 Module Nyquist impedance response under microcracks

Chemically, this means that there is a decrease in electron-hole recombination, thus increasing the width of the depletion zone; because the width is inversely proportional to capacitance size, the diffusion capacitance increases, thus explaining the larger capacitance during microcracks. The increasing series resistance with increasing irradiance can be partly attributed to the heat generated as a larger volume of majority carriers tunnels through the fractured contacts.

5.7 Chapter conclusions

In this chapter, the system's experimental design considerations and results were presented. The considerations were undertaken in line with requirements for the successful execution of linear impedance spectroscopy on a solar module through the power optimizer. The prototype's components were selected and SiC MOSFET configured for high-frequency performance. The data acquisition system was chosen and analysed for impedance estimation. The major takeaway from the acquisition system, was the use of a DAQ based method for the DC signal rejection, instead of a high-pass filter to improve acquisition time and bandwidth capability. The solar simulating light board's capabilities were assessed for better definition of the system's major controlled variables, the temperature and irradiance. The module's optimal amplitude range was determined using the FRA and from this, the module's optimal excitation frequency range was determined. Because of the module's I/V curve characteristics around maximum power, the optimal point of injection, in relation to maximum power, was determined to ensure linearity during impedance measurements.

The relevant control system's specifications were adjusted to meet the practical limitations; as such the power stage results showed good steady state results, important in impedance estimation. As mentioned during the analysis, the power stage efficiency was relatively lower than similar power stage efficiencies achieved in similar research. The major contributors are the losses along the cabling and the inductors. The switching frequency losses are another factor, specifically because it is higher than the frequencies adopted in similar work. The multiphase SMC is preferred under rated and heavier loads while the single-phase SMC is preferred under lighter load conditions for reasons discussed during the experimental analysis. The transient behaviour was, however, slower than that from the simulated results – as expected from the inertial limitations of the components. The load and input voltage disturbance tests showed that, a deviation from the nominal plant impacts both the transient and steady state behaviour of the system – a core limitation in the control architecture and strategy. This limitation was perhaps most impactful on the induced voltage excitation signals, as nonlinear behaviour was observed under variable point operation. An increase in the SNR was also another consequence of the variable operating points. It is worth noting that, although the control designed in Chapter 4 can achieve good excitation bandwidths, it remains only valid for a single operating point – one of the major reasons the perturbation signal were designed small enough such that the system remained linear; this ensured that the control system could meet its objectives. Both chirp and PRBS induced excitation signals were affected by the input voltage level and load current; this showed the need for an injection system if online impedance estimation is to be implemented. Thus, with only the power regulation and MPPT control systems, it was evident that signal injection control is required as a 3rd level to ensure good impedance measurements under all expected operational conditions.

The impedance measurements taken using the FRA showed that depending on where on I/V curve the system operates, impedance of the solar module varies. Irradiance impacts the module's diffusion parameters significantly compared to the series resistance, whereas temperature had the most significant impact of the series resistance. Based on the results, a varying operating point impacts the impedance such that the higher the module's voltage bias, the lower the impedance – in fact, all impedance parameter values decrease with increasing voltage bias. From a dynamics perspective, operating the module at lower voltages, at lower irradiances, or at high temperatures will present highest variations in the system model. Therefore, as hypothesized, the solar module's impedance contains information about the module's condition of operation; hence, characterization is possible through its impedance. The other major takeaways from the implementation are that variable point control is required for signal injection under changing operating point and load. As observed from the experimental, as well as simulation results, when injecting multiple frequencies at a time, the frequency response of the converter and module increases the crest factor of the induced excitation signal such that low frequencies are amplified, while low frequencies are attenuated. These were among the major concerns regarding implementing online impedance measurements on the solar module, as hypothesized in the introduction.

5.8 References

- [1] A. Amin, V. Kakkar, and M. Suharwerdi, "Development of low power buck convertor for enhanced light load efficiency," 2015 Annual IEEE India Conference (INDICON), 2015, pp. 1-6, doi: 10.1109/INDICON.2015.7443659.
- [2] S. N. Soheli, G. Sarowar, M. A. Hoque and M. S. Hasan, "A DC-DC Step Down Converter for Unprecedented Higher Efficiency," 2018 International Conference on Advancement in Electrical and Electronic Engineering (ICAEEEE), 2018, pp. 1-4, doi: 10.1109/ICAEEEE.2018.8643013.
- [3] M. Narimani and G. Moschopoulos, "An Investigation on the Novel Use of High-Power Three-Level Converter Topologies to Improve Light-Load Efficiency in Low Power DC/DC Full-Bridge Converters," in IEEE Transactions on Industrial Electronics, vol. 61, no. 10, pp. 5690-5692, Oct. 2014, doi: 10.1109/TIE.2014.2300063.
- [4] Noboru Katayama, Suguru Osawa, Shunya Matsumoto, Takuma Nakano, Mutsumi Sugiyama, "Degradation and fault diagnosis of photovoltaic cells using impedance spectroscopy", Solar Energy Materials and Solar Cells, Volume 194, 2019, Pages 130-136, ISSN 0927-0248.
- [5] S. Osawa, T. Nakano, S. Matsumoto, N. Katayama, Y. Saka, and H. Sato, "Fault diagnosis of photovoltaic modules using AC impedance spectroscopy," 2016 IEEE International Conference on Renewable Energy Research and Applications (ICRERA), 2016, pp. 210-215, doi: 10.1109/ICRERA.2016.7884539.
- [6] R. A. Guejia-Burbano, G. Petrone and M. Piliouline, "Impedance Spectroscopy for Diagnosis of Photovoltaic Modules Under Outdoor Conditions," in IEEE Journal of Photovoltaics, 2022, doi: 10.1109/JPHOTOV.2022.3195003.

Chapter 6

Rapid online solar module impedance estimation through SMC-based robust variable-point control

6.1 Introduction

During the study introduction several literature related to control systems implemented within linear impedance estimation was reviewed; the hypothesis made was that during variable operating conditions, the module's impedance parameters as well the converter dynamic system would impact the induced excitation signals, and hence, system linearity – thus requiring a 3-tier control system. During experimental verification, this hypothesis was accepted because, as observed from the induced excitation signals under different operating points, with only power regulation and MPPT control implemented, the magnitude of the induced excitation signals varied beyond the maximum allowable amplitude, as well as decreased below the minimum amplitude; the former led to nonlinear impedance conditions, while the latter led to high SNR and thus, nonconverging measurements.

Thus, this chapter's primary objective is the design of a variable-point excitation control system to ensure that impedance measurements can be performed on the module under variable point conditions. This is achieved by controlling the perturbation duty cycle signals within a set of parametric plants (plants with a set of variable parameters), such that the induced voltage excitation signals remain within the required linearity specifications under parametric variability in the systems operational conditions.

There are several control techniques implemented in literature related to variable-point operation, such as sliding mode [1][2], H-infinity [3][4], and QFT [5] – [8]. Sliding mode control provides the most comprehensive

robust compensation in variable-point operation; however, to achieve such capabilities, the control requires significant processing power – processing units with clock frequencies $> 100\text{MHz}$, compared to 20MHz used in this thesis. High processing power means high cost, which is not ideal for the low-cost system being pursued in this work. H-infinity works well in finding a controller gain for various operating points; however, it requires high-order controllers, which are difficult to implement practically, while requiring high processing power as well. QFT, on the other hand, offers a simpler, more cost-effective solution to variable-point control through both parametric and nonparametric uncertainties. The control technique has been shown to characterize and mitigate system structured uncertainties on converters and other applications.

In Chapter 4, small signal analysis was performed to obtain the linearized plant model of the system describing the relationship between the modulating duty cycle and the induced voltage excitation signal; this plant model showed various parameters that are subject to variations, this will be discussed in detail. With the uncertainty in these parameters addressed, quantitative feedback excitation will ensure that the excitation signals have an amplitude range that keeps the system linear, causal and time invariant for various operating points and excitation frequencies, for improved online impedance estimation.

6.2 System setup

The system setup for quantitative feedback excitation is shown in Figure 6.1 – with the QFT set up shown highlighted in red. The voltage reference, $\tilde{v}_{is,ref}(t)$ is generated within the power optimizer's algorithm. The control is performed digitally within the NI DAQ DSP unit. A multiphase boost converter was used as the excitation of choice because the inductor current ripple is not desired in this case. The converter was designed such that it regulates power delivered to battery, as well as performing impedance measurements on the solar module. To perform quantitative feedback excitation, and hence, impedance estimation on the PV module, a small voltage excitation signal was induced at the low voltage end, such that the module's output voltage, $V_{pv}(t)$:

$$V_{pv}(t) = V_{dc}(t) + \tilde{v}_{is}(t) + \tilde{v}_{sw}(t) \quad (6.1)$$

Where, $\tilde{v}_{qf}(t)$ is the induced voltage excitation signal from quantitative feedback perturbation. While the PV module's output current, $I_{pv}(t)$ is:

$$I_{pv}(t) = I_{dc}(t) + \tilde{i}_{qf}(t) + \tilde{i}_{sw}(t) \quad (6.2)$$

Where, $\tilde{i}_{qf}(t)$ is the induced current response signal from quantitative feedback perturbation. To induce the voltage excitation signal, $\tilde{v}_{in}(t)$, quantitative feedback modulating duty cycle, $\tilde{d}_{qf}(t)$ was injected into the system such that:

$$D(t) = D_{dc}(t) + \tilde{\Delta}d_{qf}(t) \quad (6.3)$$

In real systems, several parameters within the modelled system in (4.72) can change which affects the induction of the excitation signal; these are described as parametric uncertainties; where, μ_{dc} is the parametric uncertainty in the DC parameters. Equation (6.3) shows that for the PV module, there exists module parameters and the broadband excitation frequency such that, each one of them varies as shown. These variations and their impact on the induced voltage excitation signals, and subsequent impedance estimation will be analyzed in detail.

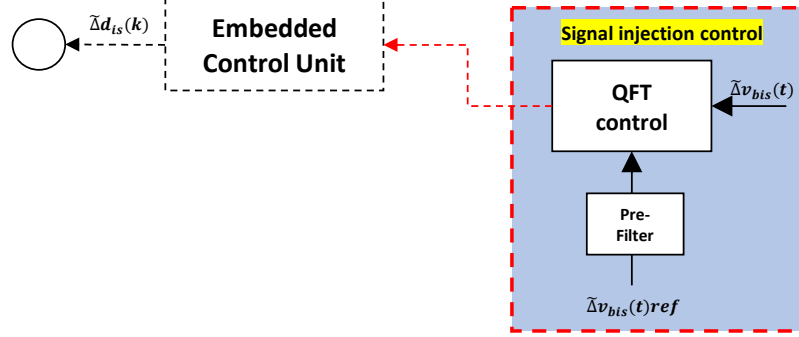


Figure 6.1 Quantitative feedback control excitation system setup

$$\forall \tilde{\Delta}z_{pv} \exists \left\{ \begin{array}{l} r_s | r_s \in \{r_{np_min}, r_{np_max}\} \Omega \\ c_{pp+} | c_{pp+} \in \{c_{pp+_min}, c_{pp+_max}\} F \\ c_{np} | c_{np} \in \{c_{np_min}, c_{np_max}\} F \\ r_{np} | r_{np} \in \{r_{np_min}, r_{np_max}\} \Omega \\ r_{pp+} | r_{pp+} \in \{r_{++_min}, r_{++_max}\} \Omega \\ \omega_{bis} | \omega_{bis} \in \{\omega_{bis_min}, \omega_{bis_max}\} Hz \end{array} \right. \quad (6.3)$$

$$\mu_{dc} \left\{ \begin{array}{l} I_L \in \{I_{Lmin}, I_{Lmax}\} A \\ \tilde{\Delta}z_{bat} \in \{\tilde{\Delta}z_{bat,min}, \tilde{\Delta}z_{bat,max}\} \Omega \\ D_{dc} \in \{D_{dc,min}, D_{dc,max}\} \end{array} \right.$$

6.3 Excitation parametric uncertainty analysis

The system in (4.72) was, thus, simulated to show how the system uncertainties affect the induced excitation signals. The induced excitation signals were then observed for various broadband perturbation frequencies, and

various module irradiance, I_r and temperature, T_k , as illustrated by Figure 6.2. In Figure 6.2 (a), the module's irradiance and temperature were varied, while the broadband excitation frequency, ω_{bis} was kept constant – the induced excitation signal was observed. The figure shows that for larger module operating voltages, the excitation signal is too large and will push the system into a non-linear range. For a lower module voltage, the excitation may be too small for module excitation and contain a large SNR. This is because under varying conditions, the impedance parameters introduce variable gain and phase to the system. Because of this, the controller designed fails to minimize the error signal and, hence produces a control action that fails to induce excitation signals within the required linear range, causing the observed phase and gain error. In Figure 6.2 (b), the module operation conditions kept the same for all frequencies, while the broadband frequency was varied linearly within the bandwidth of the module. The figure shows that the induced excitation signal amplitude varies with changing frequency, with lower frequencies affected the most in gain, phase and hence, system linearity. This results from the module's capacitive impedance – which varies non-linearly with changing frequency, which subsequently results in non-linear behaviour between the modulating injected signal and the induced excitation signal.

It is important to note that the time-averaged duty cycle, D_{dc} , inductor average current, I_L and battery impedance, R_L are also subject to significant change because of the MPPT power regulation stage – these changes were also considered. Although the inductance of any inductor varies with its average current and the increasing temperature due to that current. Figure 6.3 shows that the change in the inductance, relative to the module's operating point and operating range, is relatively negligible. The temperature coefficient of the inductors used is listed as 1% increase in the inductor's DC resistance for every 2.54 °C and a 0.0076% decrease in the inductance – which was measured using an LRC meter. As illustrated by the figure, this change is not enough to push the system out of its linear range. The decoupling and filter capacitors are also subject to variation due to changes in the system's operating temperature, as this changes the dielectric properties of the capacitors. The change in the ambient temperature is considered too low to affect dielectric properties of the device. The AC current carried by both capacitors also increases their operating temperature, but because both capacitors are designed to carry much higher frequency AC ripple currents, which are significantly lower in magnitude, the heat generated by these currents is small enough such that, the changes in both decoupling and filter capacitors' dielectric properties are considered negligible as well.

There is also a variation in the battery's charging voltage, $V_{bat}(t)$ – a voltage that is a parameter variation in the system. The injection of the modulating duty ratio thus, not only induces a voltage excitation signal at the low voltage end, but also induces a small signal AC output voltage ripple at the higher voltage end such that, the battery's charging voltage becomes:

$$V_{bat}(t) = \overline{V_{bat}}(t) + \tilde{\Delta}v_{bat}(t) + \tilde{\Delta}v_{bat,sw}(t) \quad (6.5)$$

Where, $\overline{V_{bat}}(t)$ is the stack's average charging voltage; $\tilde{\Delta}v_{bat}(t)$ is the induced small AC ripple voltage due to the perturbation duty ratio; and $\tilde{\Delta}v_{bat,sw}(t)$ is the high frequency output voltage ripple. Because $\overline{V_{bat}}(t)$ is the average DC voltage, it is designed to be constant during each of the battery's charging states. The ripple voltage, $\tilde{\Delta}v_{bat,sw}(t)$ is designed to be $\pm 2\%$ of its average value, which is within linear range of the system. However, from equations (4.56) to (4.69), the magnitude of the induced output voltage, $\tilde{\Delta}v_{bat}(t)$ can be characterized as:

$$|\tilde{\Delta}v_{bat}(t)| = \left| \frac{\tilde{\Delta}d_{in}(s) \left[-\frac{I_L}{C}s + \frac{\overline{V_{bat}}(t)(1-D_{dc}) - I_L(\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv})}{LC} \right]}{s^2 + \left(\frac{\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}}{L} + \frac{1}{\tilde{\Delta}z_{bat}C} \right)s + \left[\frac{\tilde{\Delta}z_{bat} + \tilde{\Delta}z_{pv}}{\tilde{\Delta}z_{bat}LC} + \frac{(1-D_{dc})^2}{LC} \right]} \right| \quad (6.6)$$

The induced peak response of, $\tilde{\Delta}v_c(t)$ – and hence, $V_{bat}(t)$ – are largest at the system's resonant frequency. The peak responses of the nominal operating plant, the upper and lower bounds, which include the entire family of operating points for the system, occur at frequencies that are within the bandwidth of the module. The integrity of the system's linearity at these frequencies may thus, be compromised. This peak response may also affect the MPPT algorithm because the broadband loop is much faster than the MPPT control loop, meaning that changes in the output induced voltage may result in errors in the algorithm's ability track this maximum power point. As mentioned earlier, this can easily be solved by ensuring that impedance measurements are not taken while the system is operating at maximum power – as executed by the ECU.

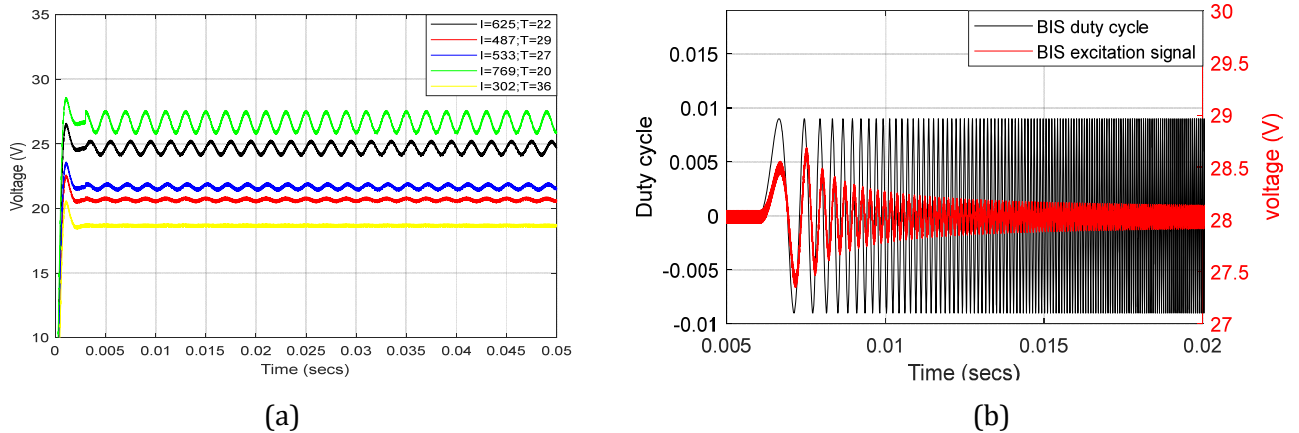


Figure 6.2 Effect of system uncertainties on the induced excitation broadband signal (a) Condition dependent parameters (b) BIS frequency

The other issue introduced by the induced voltage is the increase in the battery's charging voltage, which could lead to charging voltages that are above the maximum allowable voltages for the battery. It is worth noting, however, that these over-voltages would occur only for a short period of time while the impedance measurements are being taken. And although many batteries are designed to handle these over-voltages for a short period of time [9], it still is not desirable to allow the charging voltage to reach high levels, especially in situations

where frequent impedance analyses are to be conducted. As such, the control system was designed to reduce the peak response of the induced output voltage.

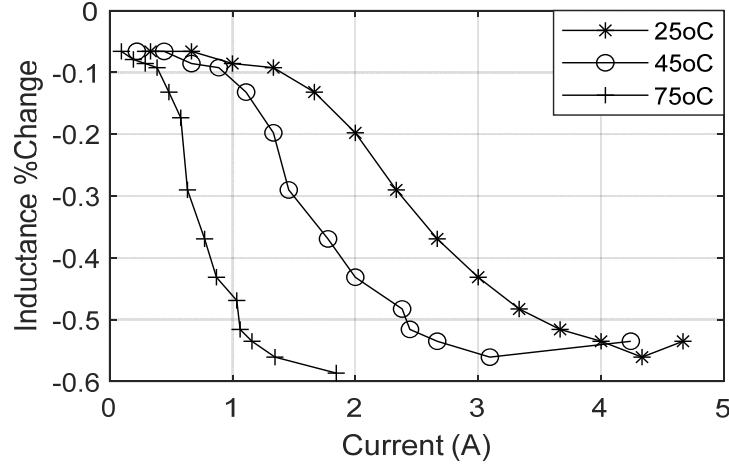


Figure 6.3 Inductance variation with average current and temperature at the switching frequency

6.4 Quantitative feedback control analysis

A QFT design approach for a LTI PV-converter impedance estimation system was undertaken, to achieve practically realizable impedance estimation in real-time, under rapidly varying conditions and system disturbances. A classical 2-degree of freedom feedback QFT system configuration was used for the synthesis, which uses feedback, controller and prefilter designs to reduce plant variations due to both parametric and nonparametric uncertainties within the PV system's operational conditions. Figure 6.4 shows that the control technique takes a classical control theory form, based on frequency domain loop shaping techniques of the open-loop plant. The characterization of the uncertainty is realized by making use of gain and phase variations evaluated on a Nichols chart, as will be discussed. The QFT controller is designed such that the system meets its desired stability, disturbance rejection and robustness; while the prefilter is designed to ensure that the system's response lies within the tracking specifications.

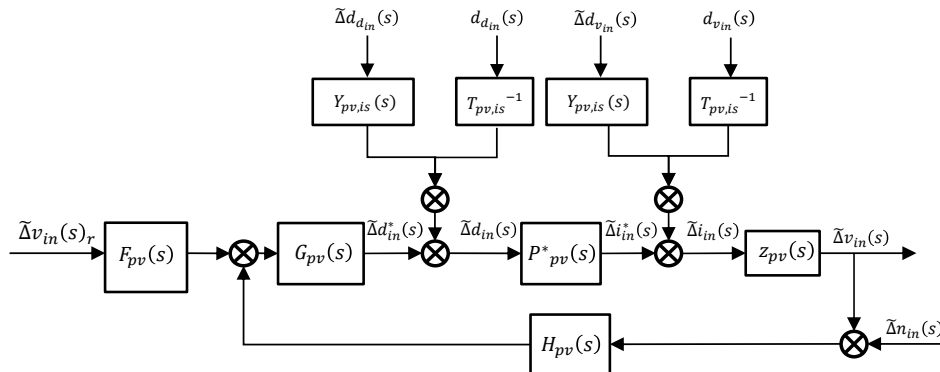


Figure 6.4 QFT control system 2-degree of freedom closed loop for voltage excitation signal induction

Where, $F_{pv}(s)$ is the prefilter; $G_{pv}(s)$ is the controller type; $P_{pv}^*(s)$ is the current loop plant; $z_{pv}(s)$ is the PV module's small AC impedance; $H_{pv}(s)$ is the feedback gain; $\Delta d_{in}(s)$ is the BIS modulating duty cycle; $\Delta i_{in}(s)$ is the induced current excitation signal; $\Delta v_{in}(s)$ is the induced voltage excitation signal. With each of the uncertainties described, a plant is defined; as such, there exist multiple gain and phase profiles that, when plotted on a Nichols plot, form a connected region referred to plant templates [10]. The size of the template is indicative of the system's uncertainty-set size. The templates are then subsequently used to define the system bounds – regions in the frequency domain where the open-loop must lie to satisfy the conditions for all plants within the set. There are 3 types of bounds: stability, performance (tracking), and sensitivity (disturbance) bounds [11]. From the set of plants within the system's uncertainty, a nominal plant is selected, and the QFT controller thus, designed such that the nominal plant meets the specifications set for each bound; loop shaping techniques are employed for this. The sensitivity and performance of the design is achieved by ensuring that the nominal plant's loop gain is above their respective bounds, whereas the system's stability is achieved by ensuring that the nominal plant lies outside the stability bounds. Although most of the design is executed on a Nichols plot, the prefilter design is undertaken on a Bode plot to satisfy the system's tracking specification.

6.4.1 Output disturbance rejection

The output disturbances in the case of impedance estimation are disturbances to the module's induced excitation voltage and hence, the converter's induced AC input voltage. For a given set of plants, $\forall P_{pv}^*(j\omega) \in \{P_{pv}^*(j\omega)\}$, the transfer function from the disturbance to the excitation voltage, $d_{vin}(j\omega)$, to the excitation voltage, $\tilde{\Delta}v_{in}(j\omega)$ is bounded such that:

$$\left| \frac{\tilde{\Delta}v_{in}(j\omega)}{d_{vin}(j\omega)} \right| = \left| \frac{G_{d_{vin}}(j\omega)}{1 + L(j\omega)} \right| \leq \beta_{d_o}(j\omega) \quad (6.7)$$

Where, $G_{d_{vin}}(j\omega)$ is the frequency response of the output disturbance; $\beta_{d_o}(j\omega)$ is the output disturbance peak response; and $L(j\omega)$ is loop transfer function, $\{H_{pv}G_{pv}P_{pv}^*\}(j\omega)$.

6.4.2 Input disturbance rejection

The concept of implementing impedance spectroscopy through a SMC is equivalent to introducing an input AC disturbance to the system. The converter operates a certain time-averaged duty cycle value; a small AC impedance spectroscopy perturbation duty cycle is then added to, effectively, “disturb” the system and monitor its steady state response to the disturbance i.e., impedance estimation. However, as observed from section 4.9.7 iii, the transients of the disturbance can cause nonlinearities in the process, resulting in reduced excitation frequencies, particularly for nonstationary parametric excitation signals. Thus, the system's response to the input disturbance for a set of plants, $\forall P_{pv}^*(j\omega) \in \{P_{pv}^*(j\omega)\}$ must be designed such that:

$$\left| \frac{\tilde{\Delta}d_{in}(j\omega)}{d_{in}(j\omega)} \right| = \left| \frac{P_{pv}^*(j\omega)}{1 + L(j\omega)} \right| \leq \beta_{d_i}(j\omega) \quad (6.8)$$

Where, $\beta_{d_i}(j\omega)$ is the input disturbance peak response.

6.4.3 Noise rejection

As discussed in the previous chapter, measurement noise is undesirable because of its impact on data acquisition; as such, for a set of plants, $\forall P_{pv}^*(j\omega) \in \{P_{pv}^*(j\omega)\}$:

$$\left| \frac{\tilde{\Delta}d_{in}(j\omega)}{d_{in}(j\omega)} \right| = \left| \frac{P_{pv}^*G_{pv}(j\omega)}{1 + L(j\omega)} \right| \leq \beta_n(j\omega) \quad (6.9)$$

Where, $\beta_n(j\omega)$ is the measurement noise peak response.

6.4.4 Tracking specification

The tracking specification defines the maximum allowable range that the closed loop system can have due to plant's uncertainties and system disturbances such that, for a set of plants, $\forall P_{pv}^*(j\omega) \in \{P_{pv}^*(j\omega)\}$,

$$T_L(j\omega) \leq F_{pv}(j\omega) \frac{P_{pv}^*G_{pv}(j\omega)}{1 + L(j\omega)} \leq T_U(j\omega) \quad \forall D_\infty \quad (6.10)$$

Where, $T_L(j\omega)$ is the lower tracking bound; and $T_U(j\omega)$ is the upper tracking bound.

6.4.5 Robust stability

To ensure stability, the following requirement on the peak magnitude of the closed loop frequency response, was used.

$$\left| \frac{L(j\omega)}{1 + L(j\omega)} \right| \leq \beta_{rs}(j\omega) \quad (6.11)$$

Where, $\beta_{rs}(j\omega)$ is the peak response of the closed loop. This constraint ensures a minimum gain margin, GM and phase margin, PM such that,

$$GM \geq 20 \log \left\{ \frac{\beta_{rs}(j\omega) + 1}{\beta_{rs}(j\omega)} \right\} dB \quad (6.12)$$

$$PM \geq 20 \sin^{-1} \left\{ \frac{1}{2\beta_{rs}(j\omega)} \right\} \text{deg} \quad (6.13)$$

6.5 Quantitative feedback control synthesis

Based on the analysis undertaken in the previous section, the controller was synthesized based on specifications to be discussed, to achieve the optimal variable point robust performance.

6.5.1 Design specifications

The design specifications as detailed in Table 4.3 are specific to a single operating. In this section, the specifications required to satisfy a wide range of operating points are details. A compensator, $G_q(s)$ and prefilter, $F_q(s)$ were designed to satisfy the following conditions:

Table 6.1 QFT design specifications

<i>Specification type</i>	<i>Specification</i>
<i>Tracking steady state error</i>	8.51% – 13.45% of $\overline{V_{pv}}$
<i>Maximum overshoot</i>	< 13.45%
<i>Settling time</i>	25ms
<i>CICM closed loop peak response</i>	5.26dB
<i>Output disturbance peak response</i>	6.44dB
<i>Input disturbance peak response</i>	6dB
<i>Noise peak response</i>	3.51dB
<i>Closed loop phase margin</i>	61.55°C

The tracking steady state error lower and upper bounds were based on the minimum and maximum excitation amplitudes, respectively; required to keep the system within its linear range. The maximum overshoot was chosen such that the excitation signal does not exceed the maximum allowable excitation amplitude. The settling time chosen based on the minimum excitation frequency of, 10Hz used in the experimental verification. The lowest excitation period is, 100ms, thus, the transients settling in less than a quarter of the period enables adequate frequency characterization.

Table 6.2 Nominal plant model and its parametric uncertainty changes

<i>Parametric uncertainty</i>	<i>Nominal value</i>	<i>Expected variation</i>
r_s	1.12Ω	[0.43,2.3]Ω
c_{pp+}	300mF	[160,500]mF
c_{np}	1mF	[0.7,130]mF

<i>Parametric uncertainty</i>	<i>Nominal value</i>	<i>Expected variation</i>
r_{np}	230 Ω	[6,575] Ω
r_{pp+}	1m Ω	[0.41,37]m Ω
ω_{bis}	28kHz	[10Hz, 100kHz]
I_{La}	3A	$\pm 58\%$
$\tilde{\Delta}z_{bat}$	24 Ω	[14,52] Ω
$\tilde{\Delta}v_{bat}$	0V	$\pm 9.88V$
D_{dc}	0.5	[0.16,0.80]

The phase margin was a consequence of the desired system overshoot, as dictated to by the equation, $PM = 75 - \%OS$. Equation (6.3) introduced the expected parameter changes, as such, Table 6.2 evaluate these changes. The uncertainty in the PV parameters was determined, by first operating the PV module under low light and warm conditions and then operating the module under fully illuminated and cooler conditions – as discussed in the previous chapter. The upper and lower bounds of the excitation frequency were determined by the excitation signal bandwidth – as determined in the previous chapter. The nominal values for the PV module parameters were chosen based on quasi-log midpoint excitation frequency of 28kHz, under operational conditions at an insolation level of 450 W/m² and 30°C.

6.5.2 Choice of frequency array

To compute the system templates and bounds, an appropriate frequency band must be chosen. The frequencies were selected by using an iterative procedure to determine the frequencies that resulted in the most significant template variations within the PV impedance bandwidth.

6.5.3 Choice of nominal plant

The nominal plant was chosen based on PV impedance parameters that would be obtained in average temperature, irradiance conditions in summer. A midpoint excitation frequency was chosen based on a quasi-log frequency distribution for the chosen excitation frequencies within the module's bandwidth.

6.5.4 Plant template generation

The templates were plotted using MATLAB on a *NICHOLS CHART* using the frequency array selected and then calculating $P_{pv}[j\omega]$ over the family of system plants – these are shown in Figure 6.5 (a). Based on the template variation, the selected frequencies would provide the necessary system robust performance to ensure the induction of a voltage excitation signal that keeps the PV system linear for various operating conditions.

6.5.5 Full system design

Based on the specifications and the analysis already undertaken the controller was synthesized as follows:

i. Stability constraint

Based on the specification, the peak response of the closed loop was kept below 1.8; this also helps keep the converter operating in CICM during the impedance measurements. The stability bounds for the system are shown in Figure 6.5 (e).

ii. Sensitivity and disturbance rejection

As discussed earlier, the system can be subjected to several disturbances at both input and output that can affect the accuracy of the measured impedance.; thus, the system must be designed to achieve certain sensitivity if system disturbances. The sensitivity function of the PV system can be defined as:

$$S_{p_{pv}}^T = \frac{1}{1 + L(j\omega)} \quad (6.14)$$

Because of PV module's wide bandwidth, the QFT controller is to be designed such that the loop gain is sufficiently large i.e., $|L(j\omega)| \gg 1$. As such, the sensitivity function can be designed so that the closed loop system can be sufficiently approximate to the prefilter, despite the PV system's parametric uncertainty. This will also significantly reduce the effect from the PV system's output disturbances. The PV-converter system was designed to be LTI and from (4.72), the system is a minimum phase plant – as such, Bode's integral theorem states that:

$$\int_0^{\infty} \log \left| \frac{1}{1 + L(j\omega)} \right| d\omega = 0 \quad (6.15)$$

For the PV system, (6.15) implies that the sensitivity function, $S_{p_{pv}}^T < 1$ up to some crossover frequency. Bode's integral theorem then implies that introducing the feedback loop, shifts the loops high sensitivity region from lower frequencies to higher frequencies when the loop gain is less than unity. Hence, for output disturbance attenuation, the sensitivity function is constraints such that:

$$\frac{1}{1 + L(j\omega)} \leq \left| \frac{1}{W(j\omega)} \right| \forall P_{pv}(j\omega) \quad (6.16)$$

Where, $W(j\omega)$ is a designed weighting function such that:

$$\left| \frac{D(j\omega)}{1 + L(j\omega)} \right| < 1 \forall \omega_{is} \quad (6.17)$$

The minimum frequency for the weighting function was selected to be lowest impedance frequency for PV module. The steady-state tracking error was selected based on the maximum amplitude of the voltage excitation signal to keep the system linear during impedance measurements. The weighting function was selected to be:

$$W(s) = \frac{0.5369(s^2 + 2.4 \times 10^2 s + 9.23 \times 10^6)}{s^2 + 1.69 \times 10^3 s + 8.36 \times 10^5} \quad (6.18)$$

iii. **Tracking bounds**

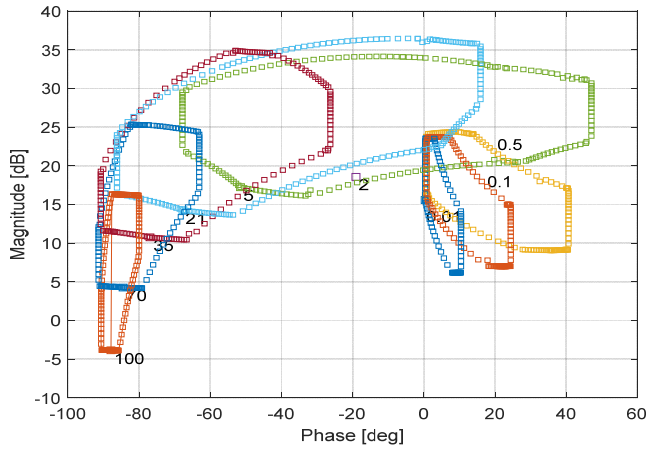
As per the specifications the tracking bounds were designed. To ensure that the response lies within the tracking bounds, a pre-filter was designed. The bound equations are as follows:

$$\left. \begin{aligned} T_L(j\omega) &= \frac{0.5369 \times 10^6}{s^2 + 7.23 \times 10^4 s + 2.88 \times 10^4} \\ T_U(j\omega) &= \frac{2.142(s + 4.72 \times 10^5)}{s^2 + 5.01 \times 10^3 s + 4.58 \times 10^6} \end{aligned} \right\} \quad (6.19)$$

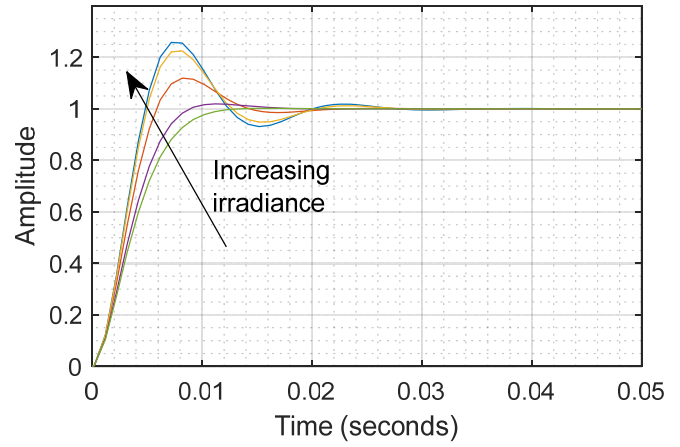
A pre-filter, $F(s)$ was designed to ensure that the response lies within the tracking bounds.

$$F(s) = \frac{0.5369}{s^2 + 5.01 \times 10^5 s + 1.99 \times 10^7} \quad (6.20)$$

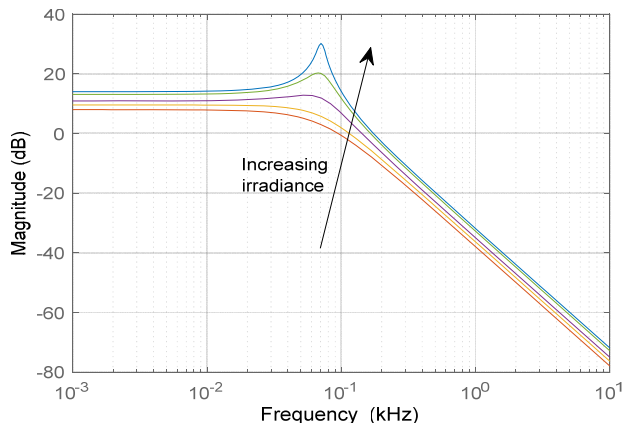
To ensure a fast response, the loop gain cross-over frequency, f_c was designed to be as large as possible and such that, $(f_s/10 < f_c < f_s/5)$; where f_s is the converter switching frequency. The appropriate poles and zeros were added to the nominal plant such that, the necessary gain could be designed to meet the required specifications. Figure 6.5 (f) shows the open loop nominal plant with the compensator. To implement the controller using digital signal processing, it was discretized using the backward Euler approximation, with a sampling period that was chosen to be twice the maximum bandwidth of the module. The controller designed takes the shape of a 2nd order lead-lag compensator – this was designed to increase the bandwidth of the system to reduce high frequency impedance attenuation, while reducing the steady state error at higher frequencies. Pole-zero pairs of the compensator were designed to add phase at, 5KHz. This ensured a high cross-over frequency and hence, increased bandwidth for the system. The resulting bandwidth was, 51kHz. This ensures a fast loop response and a bandwidth that is high enough to ensure that the converter can track the induced excitation signal, without amplifying system noise. The feedback sensor transfer function, $H_{pv}(s)$ has a 50MHz bandwidth low pass response which ensures a fast loop and accurate tracking of the signals with negligible phase error introduced at the maximum impedance frequency.



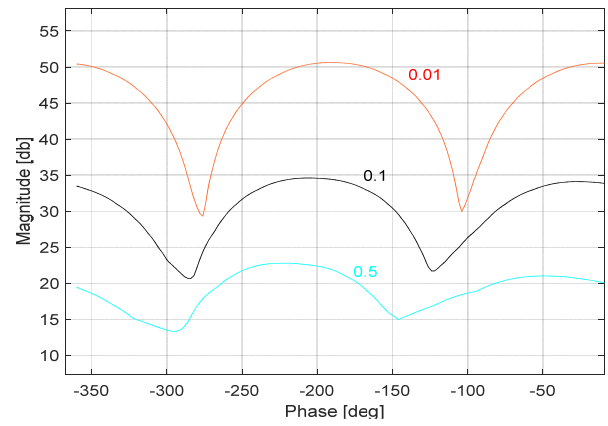
(a)



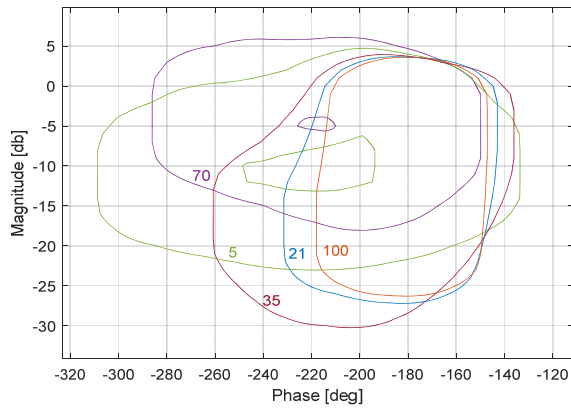
(b)



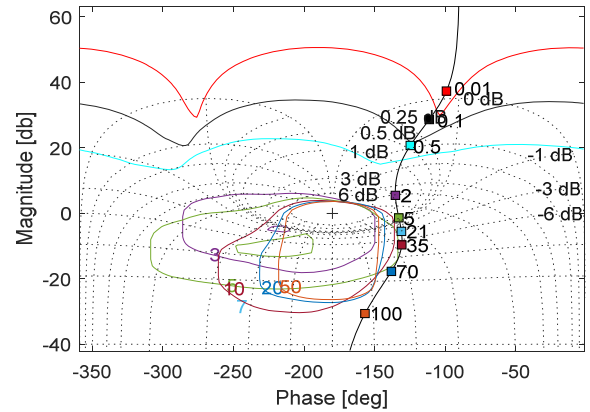
(c)



(d)



(e)



(f)

Figure 6.5 QFT synthesis (a) Templates (b) Time domain tracking specification (c) Frequency domain tracking specification (d) Tracking bounds (e) Stability and sensitivity bounds (f) Intersectional bounds with control law achieved through loop shaping

6.6 Experimental results and discussions

Earlier in the chapter, the effect of the converter and module parametric uncertainties on the excitation signals, was analysed. The results show how the integration of variable-point control helps mitigate these effects by demonstrating the induced excitation signals and impedance measurement results.

6.6.1 Induced voltage excitation signal results during variable-point conditions

In section 5.6.6, the induction of voltage excitation signals, under variable load and module voltage, were analysed in conjunction with the requirements specified in Table 4.4. For reasons already discussed, under variable conditions, the excitation signals failed to meet some of the crucial specifications like the excitation amplitude range, crest factor and excitation frequency; with quantitative feedback excitation employed, the excitation signals were once again analysed.

When either the irradiance or temperature change, the module's voltage varies. The load may also vary under certain operational conditions; thus, excitation signals during changes in both module voltage and load were assessed. Chirp excitation was used as the excitation signal of choice for the illustration. Figure 6.6 assess the control's ability to handle both output and noise rejection, tracking and robust stability specifications, while Figure 6.7 assess the control's output and noise rejections as well the control robust stability under load changes. Figure 6.6 shows that induced voltage excitation amplitude is consistent within low and high frequencies. The control achieves this by allowing the control action i.e., perturbation duty cycle amplitude, to vary such that frequency induced nonlinearities can be mitigated. The robust stability is met in both operating points within the desired specifications. The disturbance rejection at the input achieves a peak response of $1.13dB$, while the SNRs have improved to $78dB$ and $76dB$ for the $15V$ and $10V$ commands, respectively – an improvement when compared to Figure 5.18; this means that the control achieves its desired disturbance objectives. Based on the steady-state results, the control achieves amplitudes that are, 8.57% and 8.81% for the $15V$ and $10V$ commands, respectively; these results thus show that the control meets its designed tracking specifications. It worth noting that, because of the control, the excitation signal's crest factor has improved drastically to 1.57 and 1.59 for the $15V$ and $10V$ commands, respectively. This means that impedance measurements can be performed with a greater degree of accuracy during changes in the operational conditions.

Figure 6.7 shows that the control responds within its required specifications to an output disturbance, as simulated by the load changes. The peak response matches the signal's RMS value for both load levels, which is desirable for low frequency excitation. The disturbance rejection at the output achieves a peak response of $1.56dB$, while the SNRs have improved to $71dB$ and $69dB$ for the $3.1A$ and $4.5A$ commands, respectively – an improvement when compared to Figure 5.20; this means that the control achieves its desired output disturbance objectives. Based on the steady-state results, the control achieves amplitudes that are, 11.62% and 11.12% the $3.1A$ and $4.5A$ commands, respectively. It worth noting that, because of the control, the excitation signal's crest factor

has improved drastically to 1.67 and 1.78 for both nominal and light loads, respectively. This lies just outside the desired specifications but will minimally affect the linearity of the system because both load conditions meet the excitation amplitude specification; the peak response of the signal is within the desired as well. This, again, shows that measurements can be performed during load changes with minimal impact on the accuracy.

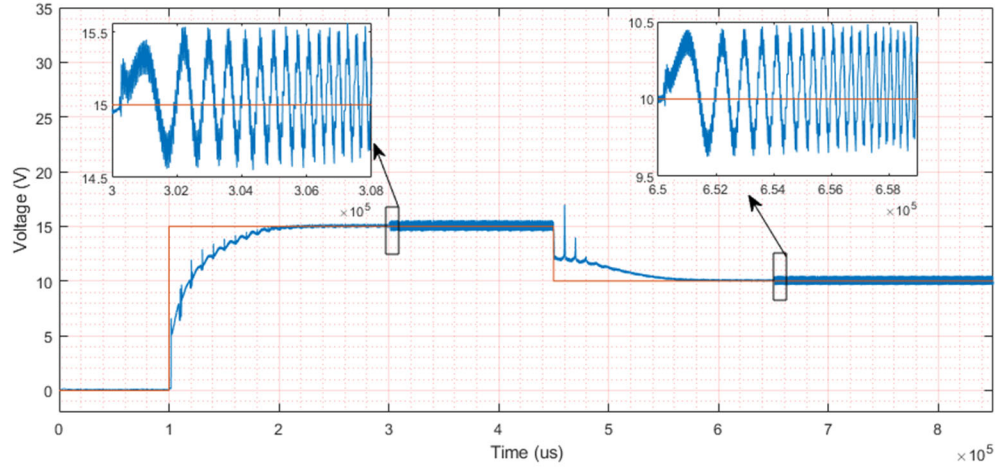


Figure 6.6 Induced voltage excitation signals during various operating points under quantitative feedback excitation, showing input and noise disturbance rejection, tracking, and robust stability

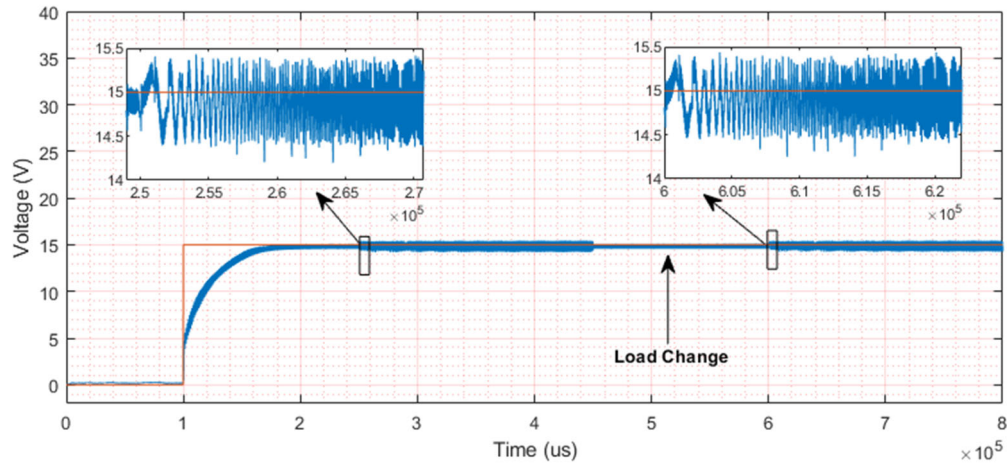


Figure 6.7 Induced voltage excitation signals during load changes under quantitative feedback excitation, showing output and noise rejection and robust stability

6.6.2 Impedance measurement results under variable conditions

The impedance measurement results are detailed in this section. The aim was to show how excitation executed through quantitative feedback control improves the impedance measurements in an online setting, under varying operational conditions. The impedance estimation results taken without quantitative feedback excitation were derived from induced voltage chirp excitation signals discussed in section 5.5.6 ii.

i. **Nyquist plots**

Figure 6.8 (a) shows that as the irradiance on the module varies, the impedance exhibits non-linear behavior. For excitation amplitudes within the desired range, the impedance does match up well with the FRA for all irradiance conditions. This non-linearity is significantly severe at lower frequency impedance values. This is because at these frequencies, the induced excitation signal has an amplitude that pushes the system into its non-linear range. The figure also shows that this increases with increasing irradiance. The non-linearity as the module temperature varies in Figure 6.8 (b), is not as severe as observed at varying module irradiance, since the system more sensitive to variations in irradiance. Figure 6.9 (a) shows that the QFT control system performs better than Figure 6.8 (a), with minimal error across all frequencies as the irradiance varies. This shows the ability of the control injection to induce excitation signal that keep the system within its linear range. The impedance error is larger at lower frequencies, because the perturbation signal is small in magnitude which results in a quantization error due to the limited resolution of the ADC. Figure 6.9 (b) also shows better impedance estimation because of the control systems wider operating range.

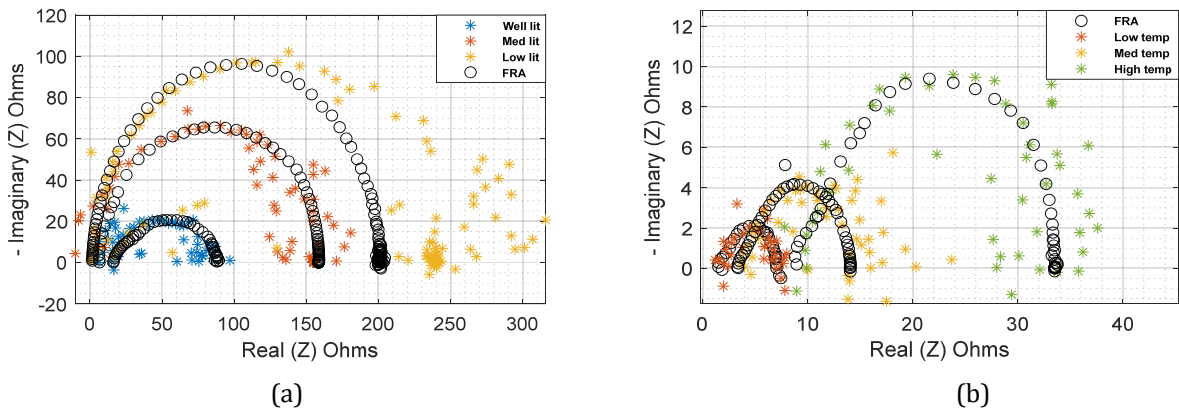


Figure 6.8 Impedance estimation on a Nyquist plot under varying conditions without QFT control injection (a) irradiance variation (b) temperature variation

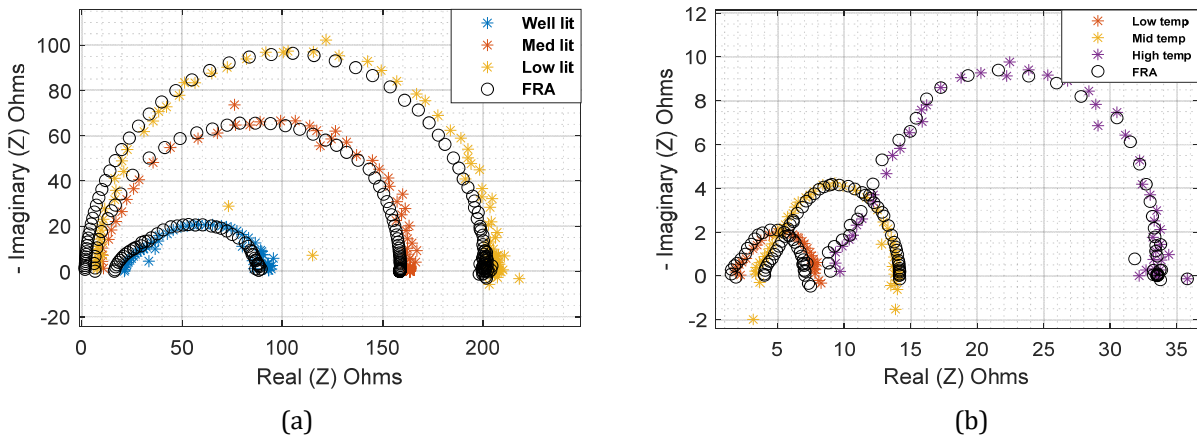


Figure 6.9 Impedance estimation shown on a Nyquist plot under varying conditions with QFT control injection (a) irradiance variation (b) temperature variation

ii. Variation of parameters

The effect of the irradiance on the resistive parameters is illustrated by Figure 6.10 (a). The irradiance decreases series and junction parameters of the module. The diffusion resistance, however, increases significantly with increasing irradiance. This significantly increases the output current produced by the module, as expected. Figure 6.10 (b) shows that the diffusion capacitance of the module increases with increasing irradiance, and it occurs due to the stored charge of minority electrons and minority holes near the depletion region. A significant irradiation on the module increases the current through the diode, and hence, a large amount of charge is accumulated near the depletion layer, leading to a large diffusion capacitance. The junction capacitance is seen to be

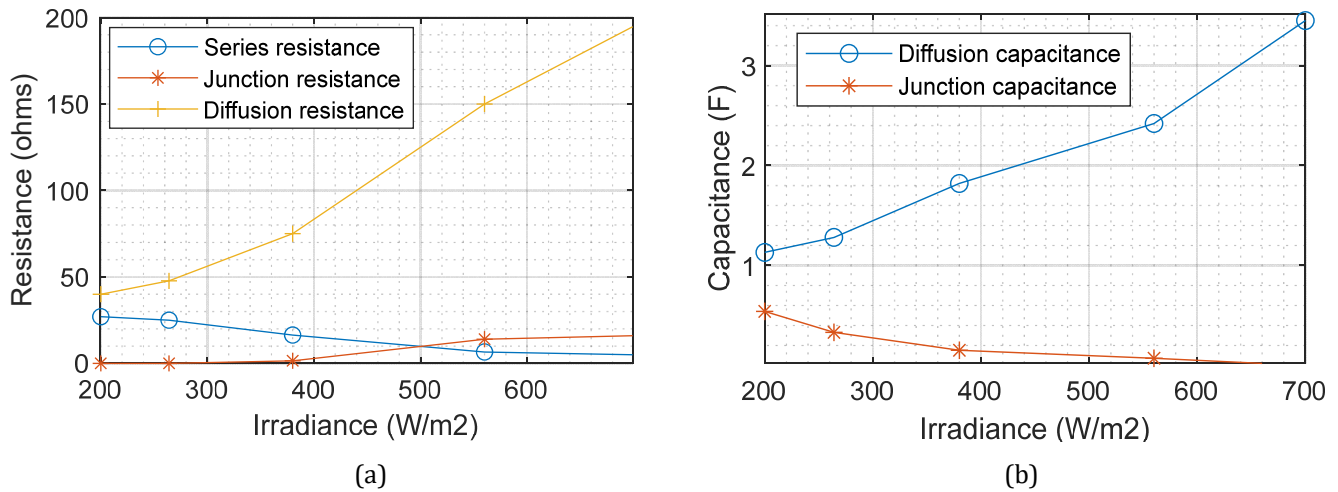


Figure 6.10 Variation of module parameters with irradiance (a) resistive (b) capacitive

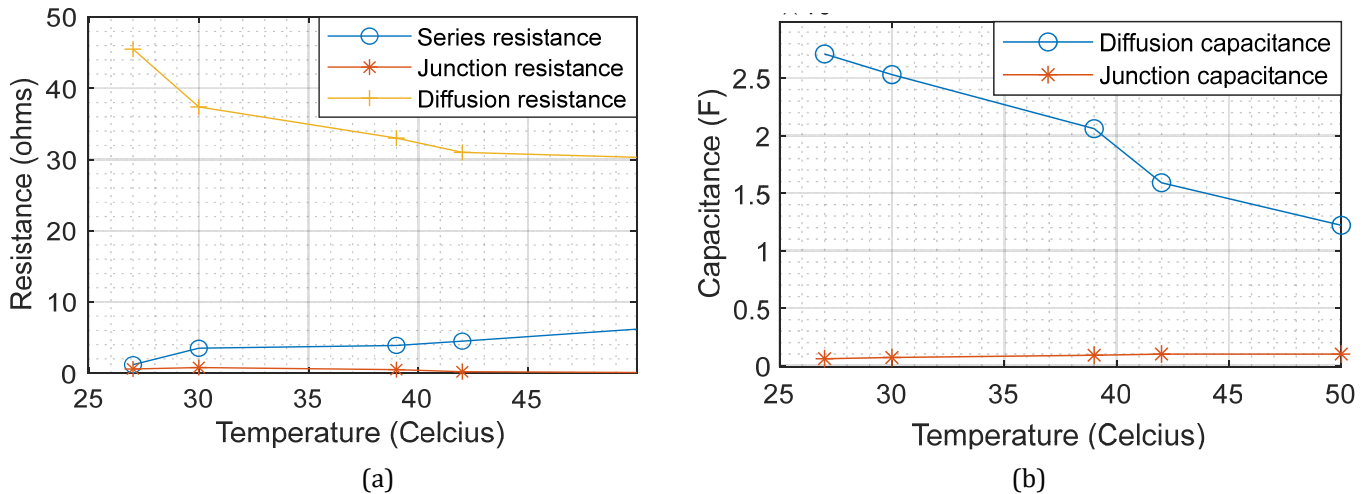


Figure 6.11 Variation of module parameters with temperature (a) resistive (b) capacitive

much smaller than the diffusion capacitance and decreases with increasing irradiance, because the increasing irradiance forward biases the module's cells. The temperature has the effect of increasing the module's series resistance as illustrated by Figure 6.11 (a) – this is due to the decreased mobility in the module's material. The

diffusion resistance, however, decreases with increasing temperature causing the open circuit voltage of the module decreasing significantly. Figure 6.11 (b) shows that an increasing temperature, increases the module's capacitive effects as expected from the behavior of capacitance under increased heat.

6.7 Chapter conclusions

In this chapter, a robust implementation of BIS of PV modules through quantitative feedback excitation, was presented. The quantitative feedback control was designed to ensure that the converter can induce voltage excitation signals that keep the system within the linear range of operation during impedance estimation. The control showed good impedance estimation when compared to an FRA under the same conditions.

When compared to the current control techniques, the proposed solution showed good impedance results that had minimal error across the PV bandwidth, particularly under rapidly varying conditions and frequency. The design also showed that the optimal excitation amplitude lies in the range of between nine and twenty percent of the module's operating voltage; within this range under lower lighting conditions, the module exhibits a first order response, but shows second order parameters at higher frequencies under high irradiance conditions.

6.8 References

- [1] G. P. Incremona, M. Rubagotti and A. Ferrara, "Sliding Mode Control of Constrained Nonlinear Systems," in *IEEE Transactions on Automatic Control*, vol. 62, no. 6, pp. 2965-2972, June 2017, doi: 10.1109/TAC.2016.2605043.
- [2] S. S. -D. Xu, C. -C. Chen and Z. -L. Wu, "Study of Nonsingular Fast Terminal Sliding-Mode Fault-Tolerant Control," in *IEEE Transactions on Industrial Electronics*, vol. 62, no. 6, pp. 3906-3913, June 2015, doi: 10.1109/TIE.2015.2399397.
- [3] Y. Zhetspisov, A. Kaibaldiyev and T. D. Do, "Robust H-Infinity Speed Control of Permanent Magnet Synchronous Motor without Load Torque Observer," 2019 IEEE Vehicle Power and Propulsion Conference (VPPC), 2019, pp. 1-4, doi: 10.1109/VPPC46532.2019.8952432.
- [4] D. N. Abara and A. Lanzon, "H-infinity-PID controller for an open-loop unstable system," 2017 IEEE 3rd International Conference on Electro-Technology for National Development (NIGERCON), 2017, pp. 134-138, doi: 10.1109/NIGERCON.2017.8281886.
- [5] S. Alsharif and M. Garcia-Sanz, "A novel multi-loop QFT robust control methodology for cascade control systems," 2015 National Aerospace and Electronics Conference (NAECON), 2015, pp. 392-397, doi: 10.1109/NAECON.2015.7443103.

- [6] Takayuki Kuwashima, Jun Imai, and Masami Konishi, "Positioning control of one link arm with parametric uncertainty using QFT method," 2008 SICE Annual Conference, 2008, pp. 3268-3271, doi: 10.1109/SICE.2008.4655229.
- [7] D. Makwana and P. S. V. Nataraj, "Automated Synstudy of Robust QFT-PID Controllers based on Piecewise Linear Approximations," 2018 IEEE 14th International Conference on Control and Automation (ICCA), 2018, pp. 787-793, doi: 10.1109/ICCA.2018.8444249.
- [8] P. Pahuja and H. M. Rai, "Design of Robust Control for Perturbed Systems Using QFT," 2012 Second International Conference on Advanced Computing & Communication Technologies, 2012, pp. 219-224, doi: 10.1109/ACCT.2012.47.
- [9] L. Held, M. Armbruster, M. Zimmerlin, M. R. Suriyah, T. Leibfried and R. Höche, "The Operation of a Battery Storage System to Avoid Overvoltages in a Low Voltage Grid," 2020 6th IEEE International Energy Conference (ENERGYCon), 2020, pp. 321-325, doi: 10.1109/ENERGYCon48941.2020.9236533.
- [10] R. Čápková, A. Kozáková, M. Minár and K. Ondřejíčka, "Robust QFT-Based Control of the DC Motor Laboratory Model," 2020 Cybernetics & Informatics (K&I), 2020, pp. 1-6, doi: 10.1109/KI48306.2020.9039892.
- [11] M. Hussain K., V. Aparna, M. S. M. Shajahan, and D. N. Jamal, "Tuning of PID Controller Using GA, SA and QFT for Three Interacting Tank Process," 2018 2nd International Conference on Trends in Electronics and Informatics (ICOEI), 2018, pp. 549-553, doi: 10.1109/ICOEI.2018.8553819.

Chapter 7

Rapid online module impedance estimation through SMC-based parametric chirp excitation

7.1 Introduction

As observed from the results during both simulation and experimental verification, induced excitation from a chirp perturbation signal, results in a signal that does not meet the required excitation specifications at both low and high excitation frequencies. This was because converters are non-linear converters, which means when implementing impedance spectroscopy – a strictly linear technique – the system must be in its pseudo-linear range. Since the crest factor in pseudo-linear excitation is required to be as close to unity as possible, the variation in the amplitude of the excitation signal, due to the system gain, is not ideal for online impedance characterization. The second issue was that these converters have at least a 2nd order frequency dynamic response – while an electrochemical device contains at least a 1st order frequency dynamic response in its AC EEC. However, the frequency response of the system describing the induction of the voltage excitation signal from a modulating duty cycle is described by these responses and hence, shows that at each operating point, the gain experienced by the excitation signal varies such that, it is largest at low frequencies and attenuates, significantly at high frequencies.

This chapter thus, demonstrates the benefit to perturbation duty cycle signal amplitude modulation in ensuring a reduced crest factor for the induced voltage excitation signal; this will help keep the system within its linear range of operation for a wider range of excitation frequencies and operating points. The modulating duty cycle signal is designed such that it contains the optimal frequency bandwidth, as well as containing an embedded amplitude-modulated profile to help mitigate the effects of the system gain on the excitation amplitude, especially at lower and higher frequencies. This will be demonstrated using an amplitude-modulated parametric chirp excitation signal, where its input parameters for the profile are designed based on the frequency response of the power converter and solar module AC EEC through frequency-domain convolution. In the time-domain, the signal's amplitude at lower frequencies would be multiplied by a lower gain profile, while at higher frequencies, multiplied by a higher gain profile. However, because the module's AC EEC parameters are frequency and operational condition dependent, the frequency response of the converter and AC EEC is required to accurately design these profiles. Hence, frequency-domain convolution is adopted, because under LTI conditions, convolution in the frequency domain is equivalent to multiplication in the time domain. An inverse Laplace transform will then be performed to obtain the amplitude profile. With this approach, the impedance error will be significantly lowered at the extreme ends of the module's bandwidth, while increasing the range of operating points within which impedance measurements can be taken in solar systems.

7.2 System setup

A multiphase configuration was chosen for the implementation. Figure 7.1 shows the system setup for amplitude-modulated chirp parametric excitation. To demonstrate the effect parametrizing the perturbation signal on the module's impedance estimation, a small voltage excitation signal was induced at the PV module, as illustrated in Figure 7.1, such that its output voltage, $V_{pv}(t)$

$$V_{pv}(t) = V_{dc}(t) + \tilde{\Delta}v_{ac}(t) + V_{sw}(t) \quad (7.1)$$

Where, $\tilde{\Delta}v_{ac}(t)$ is the induced voltage excitation signal from amplitude-modulated chirp perturbation, and output current, $I_{pv}(t)$ become:

$$I_{pv}(t) = I_{dc}(t) + \tilde{\Delta}i_{ac}(t) + I_{sw}(t) \quad (7.2)$$

Where, $\tilde{\Delta}i_{ac}(t)$ induced current response signal. To induce the voltage excitation signal, a modulating parametric excitation duty cycle signal, $\tilde{\Delta}d_{ac}(t)$ was injected into the system such that the multiphase converter's duty cycle becomes:

$$D(t) = D_{dc}(t) + \tilde{\Delta}d_{ac}(t) \quad (7.3)$$

The experimental considerations discussed in Chapter 5 were used as the underlying measures taken for the results.

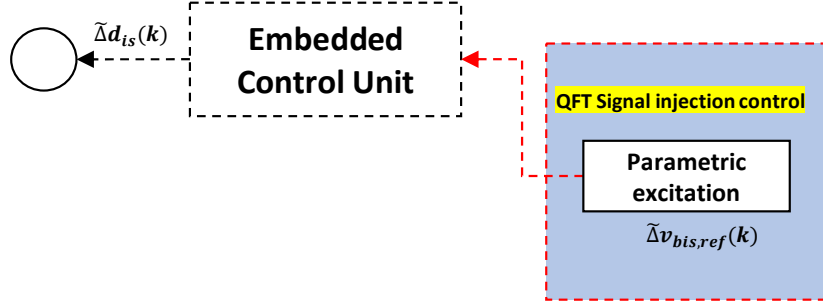


Figure 7.1 Amplitude-modulated chirp excitation system setup

7.3 Chirp excitation analysis and design

As discussed from the system setup, an excitation signal taking the form of a chirp signal was analysed and designed. This section details the mathematical definition of the chirp signal as a non-stationary signal. The wavelet transform of the non-stationary is discussed as a more precise means to analyse the time-frequency domain of the signal.

7.3.1 Chirp signal time-domain analysis

In 3.4.3 iii, the chirp signal was introduced as a sinusoidal sweep excitation signal. The signals can be classified into 3 main types, depending on the type of frequency sweep embedded within the signal, the linear, geometric, and hyperbolic chip signals [1] – [3]. For the purposes of this study, only the linear and geometric chirp excitations signals will be analysed.

7.3.2 Linear chirp

A linear chirp modulating duty ratio signal, $\tilde{d}_{cl}(t)$ can be described as follows:

$$\tilde{d}_{ac}(t) = \tilde{d}_{ac}(t) \cos(2\pi f_{is}(t)t) \quad (7.4)$$

Where, $\tilde{d}_{in}(t)$ is the modulating signal's amplitude profile; and f_{is} is the non-stationary signal's frequency profile described as:

$$f_{is}(t) = \min_a f_{is}(t) + t \left(\frac{\max_b f_{is}(t) - \min_a f_{is}(t)}{T_{is}} \right) \quad (7.5)$$

Where, $\min_a f_{is}(t)$ is the lowest signal frequency at time, $t = 0$; $\max_b f_{is}(t)$ is the signal's target frequency; and T_{is} is the signal's target time. The phase of the non-stationary signal, $\alpha_{is}(t)$ is described as the integral of frequency function of the excitation signal.

$$\alpha_{is}(t) = \min_a \alpha_{is}(t) + 2\pi \int_0^t f_{is}(\delta) d\delta \quad (7.6)$$

Where, $\min_a \alpha_{is}(t)$ is the reference phase at time, $t = 0$. Substituting (7.5) into (7.6),

$$\alpha_{is}(t) = \min_a \alpha_{is}(t) + 2\pi \int_0^t \left[\min_a f_{is}(\delta) + t \left(\frac{\max_b f_{is}(\delta) - \min_a f_{is}(\delta)}{T_{is}} \right) \right] d\delta \quad (7.7)$$

Evaluating the integral, the phase of the linear chirp excitation signal is described as:

$$\alpha_{is}(t) = \min_a \alpha_{is}(t) + 2\pi \left[\min_a f_{is}(\delta) \delta + \frac{\delta^2}{2} \left(\frac{\max_b f_{is}(\delta) - \min_a f_{is}(\delta)}{T_{is}} \right) \right] \Bigg|_0^t \quad (7.8)$$

The phase of the signal is thus:

$$\alpha_{is}(t) = \min_a \alpha_{is}(t) + 2\pi \left[\min_a f_{is}(t) t + \frac{t^2}{2} \left(\frac{\max_b f_{is}(t) - \min_a f_{is}(t)}{T_{is}} \right) \right] \quad (7.9)$$

Finally, from (7.4), (7.5), and (7.9) the time domain signal can be characterized as follows:

$$\tilde{\Delta}d_{ac}(t) = \tilde{\Delta}\widehat{d_{ac}}(t) \cos \left[\min_a \alpha_{is}(t) + 2\pi \left[\min_a f_{is}(t) t + \frac{t^2}{2} \left(\frac{\max_b f_{is}(t) - \min_a f_{is}(t)}{T_{is}} \right) \right] \right] \quad (7.10)$$

7.3.3 Geometric chirp

The geometric chirp shares a similar structure as a linear chirp, except that the frequency does not vary linearly with, but rather varies geometrically with time. The standard non-parametric modulating signal, $\tilde{\Delta}d_{cg}(t)$ can be described mathematically as:

$$\tilde{\Delta}d_{ac}(t) = \widehat{d_{ac}}(t) \cos(2\pi f_{is}(t)t) \quad (7.11)$$

Such that,

$$f_{is}(t) = \min_a f_{is}(t) A_{is} \left(\frac{\max_b f_{is}(t)}{\min_a f_{is}(t)} \right)^{T_{is}^{-1}} \quad (7.12)$$

The phase of the geometric chirp can be obtained in a similar manner to that of the linear chirp.

$$\alpha_{is}(t) = \min_a \alpha_{is}(t) + 2\pi \min_a f_{is}(t) \int_0^t A_{is} \left(\frac{\max_b f_{is}(\delta)}{\min_a f_{is}(\delta)} \right)^{T_{is}^{-1}} d\delta \quad (7.13)$$

Evaluating the integral:

$$\alpha_{is}(t) = \min_a \alpha_{is}(t) + 2\pi \min_a f_{is}(t) \left[\frac{A_{is} \left(\frac{\max_b f_{is}(\delta)}{\min_a f_{is}(\delta)} \right)^{T_{is}^{-1}} - 1}{\ln(A_{is}(t))} \right] \quad (7.14)$$

Thus, the time domain geometric chirp modulating signal can be described as:

$$\tilde{d}_{ac}(t) = \widehat{d}_{ac}(t) \cos \left[2\pi \min_a f_{is}(t) \left[\frac{A_{is} \left(\frac{\max_b f_{is}(\delta)}{\min_a f_{is}(\delta)} \right)^{T_{is}^{-1}} - 1}{\ln(A_{is}(t))} \right] \right] \quad (7.15)$$

7.3.4 Chirp signal frequency-domain analysis

To analyse the frequency response of the chirp signals, numerous frequency techniques can be employed. The Fourier transform is the most commonly used frequency domain technique; it provides frequency information in the form of frequencies and their related power [4] – [6]. The drawback to the transform is that it does not indicate the location of these frequencies in time – which does not apply to non-stationary signals. Effectively, the signal does not have the capacity to provide information for a localized signal. Therefore, a wavelet transform was considered. This transform employs multiresolution analysis, which analyses various frequencies at various resolutions. The wavelet transform of the time domain signal, $\tilde{d}_{is}(t)$, $\tilde{F}_{cg}(\beta, \gamma)$ is described below [7]:

$$\tilde{F}_{cg}(\beta, \gamma) = \frac{1}{\sqrt{|\gamma|}} \int_{-\infty}^{+\infty} \tilde{d}_{is}(t) \Psi^* \left(\frac{t - \beta}{\gamma} \right) dt \quad (7.16)$$

Where, β , is the wavelet's binary position used to translate the wavelet window function across the time signal; γ , is a scaling parameter that is the inverse of the excitation frequency, $f_{is}(t)$; and Ψ , is the wavelet. The wavelet is a small wave that serves as a basis function, as opposed to sinusoids and cosines, for the signal and acts as a window function [8]. The width and central frequency of the wavelet fully characterizes the signal [9]. An expanded wavelet is a form that can be expected at large values of the scaling parameter. This form is realized

in applications where low frequency components are more important than higher frequency components. This is because the expanded wavelet has a good low frequency resolution capacity for signals with bad time resolution. The second form a wavelet can take is the shrunk wavelet – obtained at small values for, γ . This wavelet is good at resolving high frequency components within signals with good time resolution. Thus, to perform frequency transformation on the signal, multiple dilated or compressed and shifted versions of the main wavelet, are translated across the signal. When this translated wavelet is transported across the signal, if the frequency of the wavelet matches that of the signal, the wavelet coefficients described by, $\tilde{F}_{cg}$ are high.

Based on the results in Chapter 5, it is evident that the frequency distribution of the module's Nyquist impedance response best dictates the which chirp implementation, linear or geometric, would produce the most optimal excitation results. For example, the results show that under high irradiance conditions, the module responds aggressively to higher frequency perturbations, that is, most of its impedance information is located at the higher frequency band. This means that the linear chirp will miss crucial information at lower frequencies. Although a geometric chirp would help resolve this issue, there are also 2 kinds of geometric chirp distribution, the up-chirp, and down-chirps. Because of the frequency distribution, an up-chirp is much better suited in applications where high frequency information is crucial, because of the transient behaviour during signal injection.

7.4 Amplitude modulated parametric chirp excitation analysis and design

The parametric signal was designed such that its amplitude profile could be tailored to the injected signal's frequency response. To design the input parameters for the modulating parametric signal, the series configuration of the PV module's small signal and the converter's dynamics responses were considered. The amplitude-modulated signal, $\tilde{d}_{ac}(s)$ must thus, be designed to ensure that the amplitude points are tailored to the chosen module AC EEC and converter.

Convolution is a mathematical technique that is used to determine the output a system produces for an input signal from a system with zero initial conditions. An LTI system can be completely characterized by its impulse response [10]. A continuous-time impulse signal's shifting property dictates that the input signal to a system can be fully described by the integral of scaled and shifted impulses [11]. Through the linearity property, the output signal is calculated as the sum of scaled and shifted impulse responses and, thus, as the integral of scaled and shifted impulse responses; this is the principal concept behind convolution. Convolution can, therefore, be used to determine an LTI system's output from its input and impulse response. Recall that the convolution of a signal can be computed as follows:

$$(x * y)(t) = \int_{-\infty}^{\infty} x(\tau) y(t - \tau) d\tau \quad (7.17)$$

Because the chirp signal is a parametric non-stationary signal, as well as the fact that frequency analysis is more important in impedance spectroscopy, its frequency domain characteristics are more important. From both the simulation and experimental excitation results, it is evident that the designed perturbation signal must contain smaller amplitudes at lower frequencies to mitigate the large gain experience; it must also contain larger amplitudes at higher frequencies to help improve the gain. Intuitively, the amplitude profile of the perturbation signal can be designed by proportionally inversing, in the time domain, the gain profile observed from the induced excitation signal around some critical axis, and then multiplying that signal with the original perturbation signal. However, because the frequency response of the system is well understood, and the fact that the multiplication of 2 signals in time is equivalent to the convolution of the signals in the frequency domain; thus, the amplitude-modulated parametric perturbation signal can be designed by evaluating the convolution of the original perturbation signal's wavelet transform and the system's frequency response such that,

$$\tilde{\Delta}d_{ac}(s) = \int_0^s \tilde{\Delta}F_{ac}(\tau, \gamma) * -P_{pv}(s - \tau) d\tau \quad (7.18)$$

For both the linear and geometric chirp excitation signals. It is worth noting that the result in (7.18) is possible because impedance spectroscopy makes use of small signal perturbations such that the system remains linear and time-invariant (LTI). The system described in (4.72) – derived from the small signal analysis of the PV module's AC EEC model and multiphase converter – is an LTI system as well, hence the convolution is valid. The mathematical convolution of (7.18) was computed on MATLAB, and then the inverse Laplace transform of the result used to describe the parametric amplitude-modulated perturbation chirp duty cycle signal, such that:

$$\tilde{\Delta}d_{ac}(t) = \mathcal{L}^{-1} \left\{ \int_0^s \tilde{\Delta}F_{ac}(\tau, \gamma) * -P_{pv}(s - \tau) d\tau \right\} \quad (7.19)$$

A MATLAB program was written to compute the results of both (7.18) and (7.19).

7.5 Experimental results and discussions

The characteristics of the excitation signal were graphically illustrated to show the justification of using a particular frequency spread. Both time and frequency domain results were analysed thoroughly for this. The induced excitation signals were analysed in relation to the linearity requirements, thus also showing the benefits of parametrization. Lastly, the impedance measurements of both the bode and Nyquist plots were discussed.

7.5.1 Chirp signal frequency domain results

The analyses into frequency resolutions for the linear and geometric excitation signals were undertaken. The successful execution of impedance spectroscopy requires good frequency resolution. Figure 7.2 shows the linear chirp's frequency distribution with good frequency resolution. In impedance spectroscopy, good frequency is preferred for better impedance characteristics. Figure 7.3 shows the spectrum for the geometric chirp signal with good frequency resolution also as preferred. The results in section 5.5.7 showed that the frequency spread in the module's response is significantly dependent on the operating conditions. Recalling that, under high irradiance conditions, the module responds better to higher frequency components – more specifically, on the Nyquist plot, frequencies above 20kHz dominate the module's response. As such, higher frequency points from the excitation signal are preferred over lower frequency points. The linear chirp would not be the most optimal choice under these conditions because it prioritizes a linear frequency spread, whereas the geometric up-chirp on the other hand can prioritize more higher frequency points; hence, characterization under

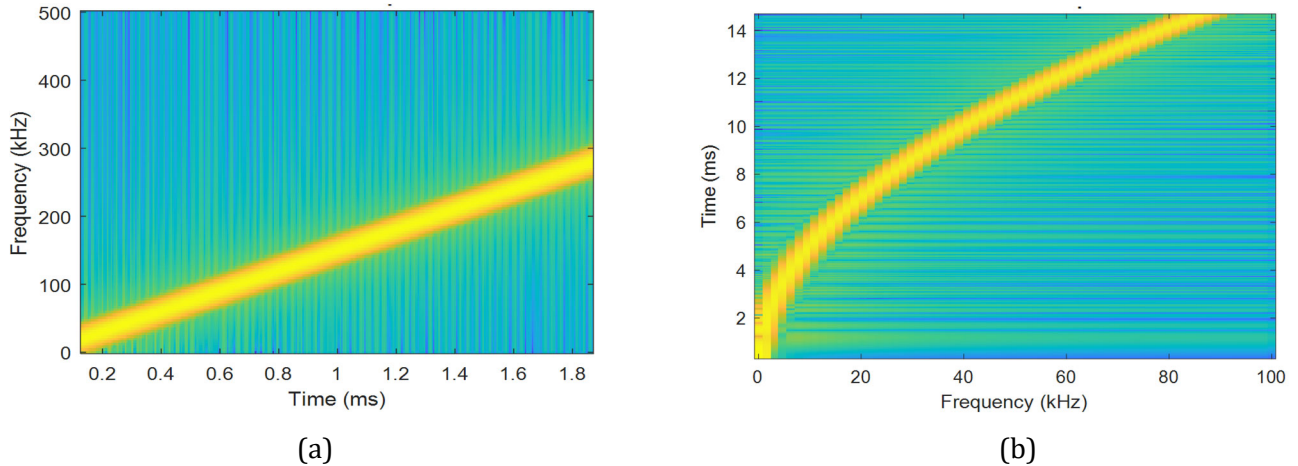


Figure 7.2 Spectrogram analysis for chirp excitation signals (a) linear and (b) geometric

high irradiance conditions – specifically insolation levels $\geq 600 \text{ W/m}^2$, based on laboratory testing – geometric up-chirp excitation for the module is preferred. Under low irradiance to dark conditions, however, lower frequency components dominate the Nyquist plot. As was the case under high irradiance conditions, the linear chirp does not provide the most optimal results. The geometric down-chirp excitation signal does provide the most optimal results as it prioritizes lower frequency components. It is worth noting that the linear chirp does apply to conditions occurring at insolation levels, $423 \text{ W/m}^2 \lesssim \text{Linear Chirp} \lesssim 536 \text{ W/m}^2$ – based on laboratory testing with the solar simulating light board. This means that under these conditions, there is a linear distribution of the frequency points in the module's response on the Nyquist plot.

7.5.2 Parametric chirp excitation signal induction analysis

Having gone through the limitations of the nonparametric signals, the discussion now turns to the idea of

parametrization. When a modulating duty ratio is used to induce a voltage excitation signal at the input of the device, the resulting excitation signal from the parametric modulating signals are illustrated in Figure 7.3.

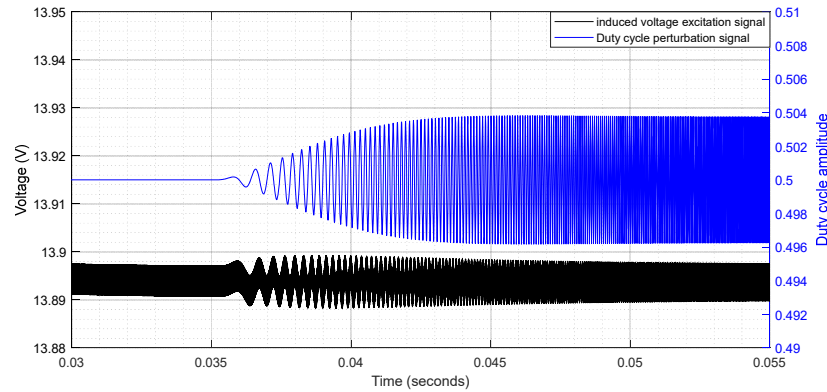


Figure 7.3 Parametric module chirp excitation

Compared to the results obtained in Figure 5.18, Figure 7.3 shows that the parametric algorithm produces a modulating duty ratio that contains lower amplitudes at lower frequency components, evidently as a mitigation measure for the effects of the system gain, as discussed in the previous section. The induced voltage excitation signal now shows a near uniform crest factor, as desired. The amplitudes at lower and higher frequencies now meet the required levels for linear excitation – which increases the optimal perturbation amplitudes.

7.5.3 Impedance estimation results

Figure 7.4 shows that nonparametric excitation exhibits nonlinear behaviour at lower frequencies as hypothesized because of reasons discussed in the previous sections, while parametrizing the excitation signal shows less nonlinear behaviour. High frequencies show phase shift which impacts the accuracy of the series resistance, this is due to the phase response of the converter at high frequencies. This phase shift is more evident under hotspot conditions where the impedance of the module is higher, as illustrated by Figure 7.5.

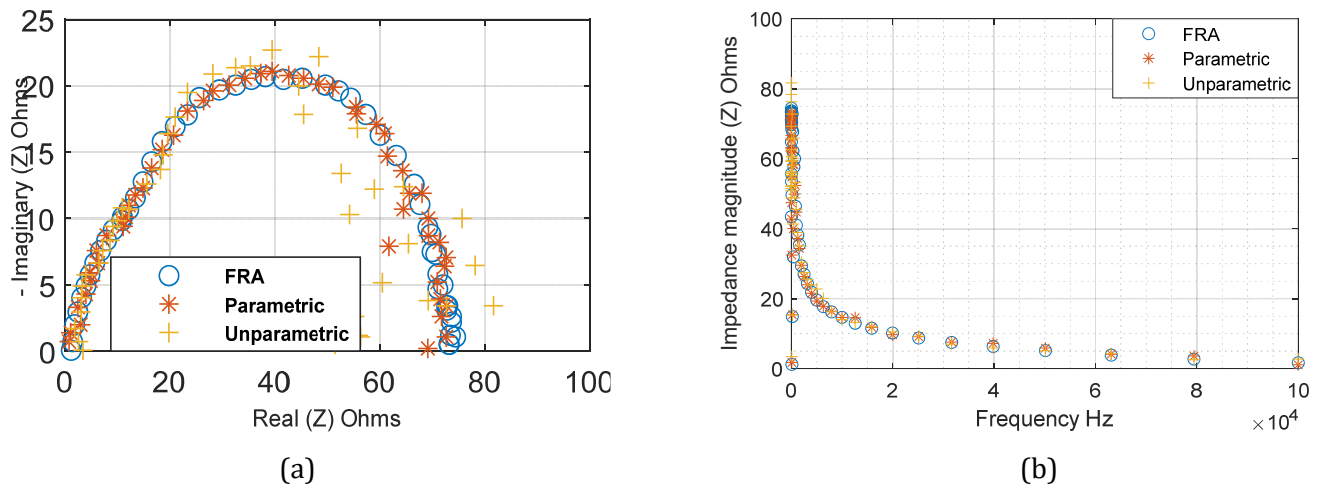
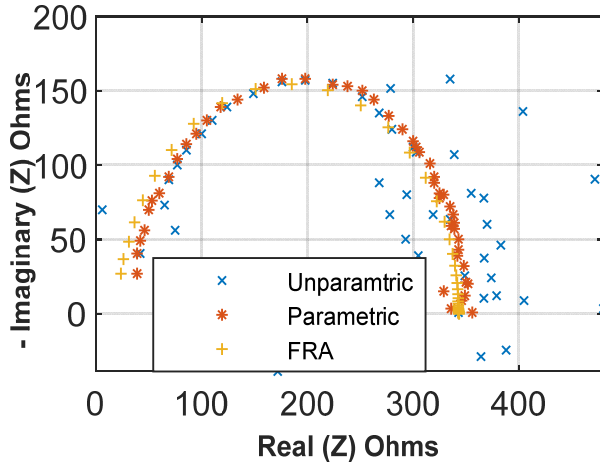
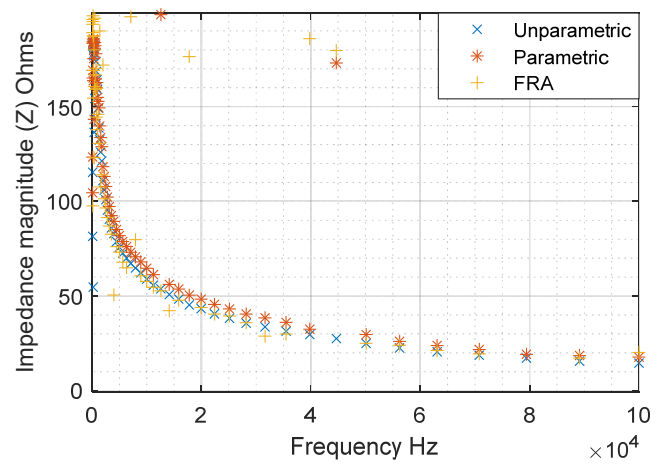


Figure 7.4 PV system impedance estimation under normal well-lit conditions (a) Nyquist plot (b) Bode plot

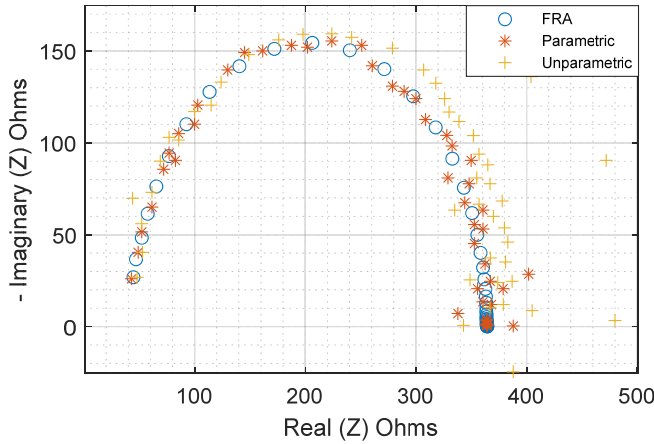


(a)

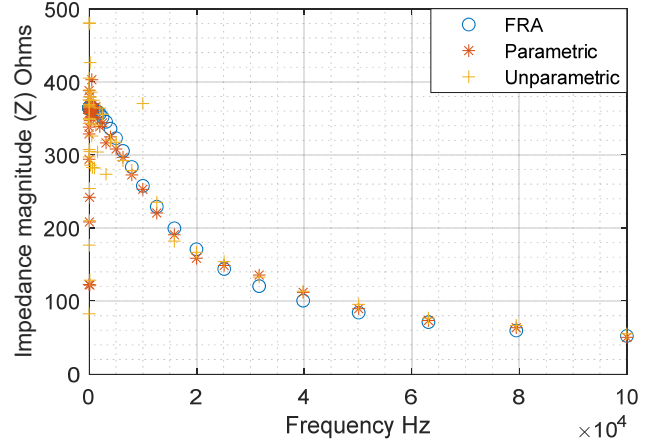


(b)

Figure 7.5 PV system impedance estimation under hotspots (a) Nyquist plot (b) Bode plot



(a)



(b)

Figure 7.6 PV system impedance estimation under partial shading conditions (a) Nyquist plot (b) Bode plot

Under hotspot conditions, the resistance and capacitance of the module increases significantly, and this is captured more effectively by the parametric excitation signal. Whereas the nonparametric signal still shows nonlinear behaviour at lower frequencies. Figure 7.6 shows that the designed system also performs better under partially shaded conditions. However, the non-parametric excitation signal shows better precision at lower frequencies, while the parametric excitation signal shows better accuracy.

7.6 Chapter conclusions

In this chapter, a parametric excitation signal design approach to broadband frequency characterization of a monocrystalline PV module through a multiphase switch-mode boost converter was undertaken. The aim of the study was to show how parametrizing the modulating duty ratio perturbation signal can mitigate the effect of the system gain on the induced voltage excitation signals, resulting from the use of a switch-mode converter as

the excitation source. By performing the frequency-domain convolution between the small signal converter characteristics and the chirp modulating duty ratio in the frequency domain, the parametric excitation signal was designed. Analysis into the type of chirp excitation signal showed that depending on the operational condition, both linear and geometric chirp signals can be used, with the linear chirp favoring conditions of medium light, and the geometric chirp favoring conditions under high irradiance and low irradiance conditions.

The results showed that although parametrization introduces complexity in the system design, it works in designing device excitation signals that keep the system in its linear range of operation during impedance characterization. The resulting impedance showed good online impedance error when compared to an FRA under hot spots conditions; the hot spots results are in line with results obtained in [12]. This enables the system to operate adequately in online conditions with a significant number of variations. It is worth noting that the chirp signal was used for battery and fuel cell impedance spectroscopy characterization in [13] and [14], respectively. However, the maximum observed bandwidth for a Li-ion $2kHz$, while for a Lead-acid battery, it is $6kHz$. The maximum observed bandwidth for a fuel cell is $10kHz$, as per the research. These loop bandwidths are generally within an SMC's loop bandwidth; hence, the effects of a broadband frequency are minimal. Therefore, this application is more suited towards wide impedance bandwidth devices.

7.7 References

- [1] O. Aldimashki and A. Serbes, "Performance of Chirp Parameter Estimation in the Fractional Fourier Domains and an Algorithm for Fast Chirp-Rate Estimation," in *IEEE Transactions on Aerospace and Electronic Systems*, vol. 56, no. 5, pp. 3685-3700, Oct. 2020, doi: 10.1109/TAES.2020.2981268.
- [2] D. Sysak, G. Jaromi and P. Biernacki, "Implementation and verification of Nonlinear chirp signals in hardware system," 2021 IEEE Region 10 Symposium (TENSYP), 2021, pp. 1-4, doi: 10.1109/TENSYP52854.2021.9550831.
- [3] C. Wang and J. Yao, "Nonlinearly chirped microwave pulse generation using a spatially discrete chirped fiber Bragg grating," 2009 International Topical Meeting on Microwave Photonics, 2009, pp. 1-4.
- [4] Soo-Chang Pei, Min-Hung Yeh and Tzyy-Liang Luo, "Fractional Fourier series expansion for finite signals and dual extension to discrete-time fractional Fourier transform," in *IEEE Transactions on Signal Processing*, vol. 47, no. 10, pp. 2883-2888, Oct. 1999, doi: 10.1109/78.790671.
- [5] W. E. Pelton, A. Stoica and N. Yossakda, "Sliding aperture Fourier transform (SAFT): a low latency continuous process," 2001 CIE International Conference on Radar Proceedings (Cat No.01TH8559), 2001, pp. 585-589, doi: 10.1109/ICR.2001.984785.

- [6] S. C. Pei, W. L. H. Sue, and J. J. Ding, "Discrete Fractional Fourier Transform Based on New Nearly Tridiagonal Commuting Matrices," in *IEEE Transactions on Signal Processing*, vol. 54, no. 10, pp. 3815-3828, Oct. 2006, doi: 10.1109/TSP.2006.879313.
- [7] S. D. Thepade and P. Bidwai, "Iris recognition using fractional coefficients of transforms, Wavelet Transforms and Hybrid Wavelet Transforms," 2013 International Conference on Control, Computing, Communication and Materials (ICCCCM), 2013, pp. 1-5, doi: 10.1109/ICCCCM.2013.6648921.
- [8] Cai-Xia Deng, Yuan-Yan Tang, and Zuo-Xian Fu, "Reproducing kernel space based on GABOR wavelet transform," 2008 International Conference on Wavelet Analysis and Pattern Recognition, 2008, pp. 644-649, doi: 10.1109/ICWAPR.2008.4635858.
- [9] H. Toda and Zhong Zhang, "Perfectly translation-invariant complex wavelet packet transforms," 2009 International Conference on Wavelet Analysis and Pattern Recognition, 2009, pp. 374-379, doi: 10.1109/ICWAPR.2009.5207462.
- [10] B. Ren and D. -y. Yu, "Evaluate the Convolution Integral and Convolution Sum Use Compact Formula," 2011 7th International Conference on Wireless Communications, Networking and Mobile Computing, 2011, pp. 1-4, doi: 10.1109/wicom.2011.6039951.
- [11] C. Li, H. K. Kwan, and X. Qin, "Revisiting Linear Convolution, Circular Convolution and Their Related Methods," 2020 13th International Congress on Image and Signal Processing, Biomedical Engineering, and Informatics (CISP-BMEI), 2020, pp. 1124-1131, doi: 10.1109/CISP-BMEI51763.2020.9263607.
- [12] S. Osawa, T. Nakano, S. Matsumoto, N. Katayama, Y. Saka, and H. Sato, "Fault diagnosis of photovoltaic modules using AC impedance spectroscopy," 2016 IEEE International Conference on Renewable Energy Research and Applications (ICRERA), 2016, pp. 210-215, doi: 10.1109/ICRERA.2016.7884539.
- [13] Olakunle Alao & Paul Barendse, "Online Condition Monitoring of Sealed Lead Acid & Lithium Nickel-Cobalt-Manganese Oxide Batteries using Broadband Impedance Spectroscopy," ECCE – IEEE, Portland – Oregon, USA, September 2018.
- [14] F.A Aroge, P.S. Barendse, "Time-Frequency Analysis of the Chirp Response for Rapid Electrochemical Impedance Estimation", IEEE Energy Conversion Congress and Expo (ECCE'18), Portland Oregon, 2018.

Chapter 8

Rapid online solar module impedance estimation through SMC-based amplitude-modulated pseudo-random binary sequence excitation

8.1 Introduction

As analysed in Chapters 4 and 5, PRBS excitation results showed poor crest factors with induced excitation signal crest factor, as well as poor high frequency excitation. As mentioned in the introduction, the use of binary broadband signals has been investigated on battery systems to further decrease measurement time and reduce systems costs, in online system identification [1] – [4]. The major issues that have not been addressed are the frequency induced non-linearities that arise from the induction of the excitation signal through a non-linear switch-mode converter. The introduction of a digital excitation signals, whose frequency is randomized over the system bandwidth, causes a large uncertainty around the induced non-linearities during impedance estimation. PV modules have non-linear behaviour over a wider frequency bandwidth – particularly around maximum power, where the system is at its most non-linear region. Because of this, the accuracy in determining the impedance would still be compromised using PRBS or ternary sequence, especially during instances of partial shading where there can be various maximum power points.

This chapter thus proposes fast digital perturbation by making use of an adaptive 2-level optimized APRBS parametric broadband excitation signal. The parametric APRBS ramp function design points were synthesized using an adaptive model-based online DoE. The input space for signal amplitude optimization was defined based on the non-linear relationship between the converter APRBS modulating

duty cycle and APRBS voltage excitation signal. The APRBS will fasten impedance estimation results, but with better frequency-induced system non-linearity mitigation and ensure accurate impedance estimation over a wide range of system operating points.

8.2 APRBS excitation system setup

To perform APRBS based BIS on the PV module, a small APRBS voltage excitation signal was induced at the PV side, such that the PV voltage, $V_{pv}(t)$ and current, $I_{pv}(t)$ may be represented as:

$$V_{pv}(t) = V_{dc}(t) + \tilde{\Delta}v_{aprb}(t) + V_{sw}(t) \quad (8.1)$$

Where, $\tilde{\Delta}v_{aprb}(t)$ is the APRBS induced voltage excitation signal.

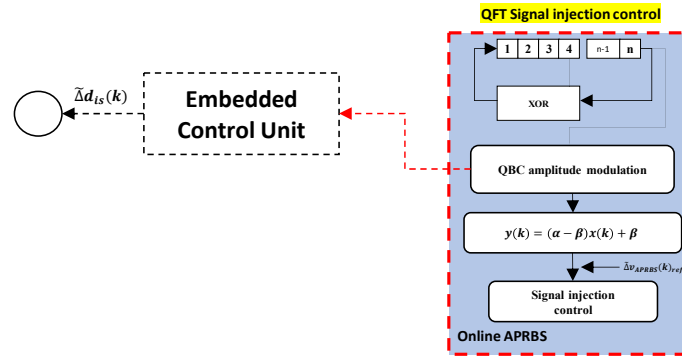


Figure 8.1 Amplitude-modulated pseudorandom binary sequence parametric excitation system setup

$$I_{pv}(t) = I_{dc}(t) + \tilde{\Delta}i_{aprb}(t) + I_{sw}(t) \quad (8.2)$$

Where, $\tilde{\Delta}i_{aprb}(t)$ is the expected APRBS response current signal. To induce the excitation signal, $\tilde{\Delta}v_{aprb}(t)$, a modulating duty ratio, $\tilde{\Delta}d_{aprb}(t)$ was injected into the system such that:

$$D(t) = D_{dc} + \tilde{\Delta}d_{aprb}(t) \quad (8.3)$$

The ECU thus, was designed to ensure that the power regulation and BIS systems work seamlessly together for better impedance estimation.

8.3 PRBS amplitude modulation design

Although the PRBS signal is ideal for linear systems under direct excitation for reasons already mentioned, for systems with induced excitation, this signal presents issues because the induction of the excitation signal contains varying frequency and operating point induced non-linearities. This can be mitigated by modulating the amplitude of the PRBS perturbation duty cycle, $\tilde{\Delta}d_{aprb}(t)$ such that lower and higher frequency regions within the signal have amplitudes that help maintain a uniform induced

excitation signal. Section 4.9.7 vi, focused on the optimization of the excitation bandwidth on the PRBS signal, those design considerations were carried over when designing the APRBS signal.

The APRBS, described fully by a series of design points, d_p and a sequence of dwell times, T_p contains the same properties of the PRBS signal. In a practical system the APRBS, defined by a series of step functions, cannot be realized because the passive components of the switch-mode converter and PV module's capacitive impedance prevent the voltage and current from sudden changes [5][6]. It is therefore desirable to design the signal by a series of N ramp functions such that characterized by design points, d_{p-1} and d_p at instances, $p - 1$ and p ; and dwell time, as illustrated by (8.4):

$$R_{APRBS} = \sum_{p=1}^N \Delta R_{APRBS}(d_p, d_{p-1}, T_{p,n}) \quad (8.4)$$

Each ramp function, ΔR_{APRBS} exists at discrete-time input sequences defined by sequence time, $T_{p,n}$ with a sampling time of, n which implies that the ramp function:

$$\Delta R_{APRBS}(d_p, d_{p-1}, T_{p,n}) = \langle \Delta r_{\Delta k+1} \rangle_{t-1}^{T_p} \quad (8.5)$$

Where:

$$\Delta k = \frac{1}{n} \sum_{x=1}^{p-1} T_x \quad (8.6)$$

And the singular ramp function:

$$\Delta r_{k+1} = \begin{cases} d_p, & \frac{T_{d,p}}{n} \leq z \leq \frac{T_p}{n} \\ s_{APRBS} * z + d_{p-1}, & 1 \leq z \leq \frac{T_{d,p}}{n} \end{cases} \quad (8.7)$$

Where, s_{APRBS} is defines the slope the ramp functions. The limit of this slope as the change in its dwell time approaches half of its value is given by:

$$s_{APRBS} = \lim_{T_{d,p} \rightarrow \Delta t} \varphi \left[\frac{s_{APRBS,max}}{T_{d,p}} \right] \quad (8.8)$$

Where, $s_{APRBS,max}$ is the maximum allowable slope with the ramp time, $T_{d,p}$ defines as:

$$T_{d,p} = \frac{1}{n} \|\varphi_x\| \quad (8.9)$$

And:

$$\varphi_x = \frac{d_p - d_{p-1}}{s_{APRBS,max}} \quad (8.10)$$

The ramp time and signal dwell time are a multiple of the sampling time and each slope instance is limited by, $s_{APRBS,x} \leq s_{APRBS,max} \forall x$. This implies that the ramp function has a length that depends on the ramp time and dwell time [7]. The APRBS excitation signal was designed such that, the relative

difference in the power spectrum density between it and the PRBS is less than 2%, by ensuring that it satisfied the condition [8]:

$$T_a \leq T_{n,min} \frac{1.02}{2\pi} \quad (8.11)$$

Where, T_a is an incremental ramp time; and $T_{n,min}$ is the smallest system time constant [9]. To ensure signal robustness, the design points input space will be design based on the dynamic system. Euler's forward approximation was used to characterize the dynamic system, as illustrated by (8.12).

$$\begin{bmatrix} i_L(k+1) \\ v_c(k+1) \end{bmatrix} = \begin{bmatrix} -(\tilde{\Delta}z_L + \tilde{\Delta}z_{pv}) & \frac{(1-d)}{L} \\ \frac{(1-d)}{C} & -\frac{1}{\tilde{\Delta}z_L C} \end{bmatrix} \begin{bmatrix} I_L(k) \\ V_c(k) \end{bmatrix} + \begin{bmatrix} \frac{1}{L} & 0 \\ 0 & \frac{1}{C} \end{bmatrix} \begin{bmatrix} V_{dc}(k) \\ I_{dc}(k) \end{bmatrix} \quad (8.12)$$

Where, $\tilde{\Delta}z_L(k)$ is the battery bank impedance. Equation (8.12) shows that the induction of the excitation signal is significantly dependent on the module parameters $\tilde{\Delta}z_{pv}(k)$, and hence, excitation frequency; and the module's operating voltage and current. This dynamic description will form the input space for the adaptive online parametric profile for the excitation signal. The design points for the signal were designed such that they were placed within the input space where discrete-time input and output committee models, in parallel configurations, are strongly disagreeing. Each committee was characterized by a sequence of, M models such that [10][11]:

$$\hat{v}_k^{(i)} = \sum_{i=1}^M \lambda_i f(\beta_k^{(i)} \gamma_k^{(i)}) \quad (8.13)$$

Where, $\hat{v}_k^{(i)}$ is the output of each committee model; λ_i are the designed signal's weighting coefficients; $\beta_k^{(i)}$ describes the parameter vector; $\gamma_k^{(i)}$ is the signal's regressor vector. The regressor vector consists of input and output models such that [12]:

$$\gamma_k^{(i)} = [\hat{v}_{k-1}^T, \dots, \hat{v}_{k-l}^T, \hat{d}_{k-1}^T, \dots, \hat{d}_{k-j}^T]^T \quad (8.14)$$

For good design points optimization, the criteria the DoE was defined under boundary conditions as Model Predictive Control (MPC). As such, the cost function to be minimized over the APRBS time horizon, T_p can be described as [13]:

$$J_k(\Delta R_{APRBS}, k, T_p) = \sum_{j=k}^{k+N_i} \sum_{i=1}^M \lambda_i M(\theta_k^{(i)} \varphi_j^{(i)}) \quad (8.15)$$

Where, $M(\theta_k^{(i)} \varphi_j^{(i)})$ are the model dependent costs such that:

$$\hat{v}_j^{(i)} \in \mathbb{I} \quad \hat{d}_j^{(i)} \in \mathbb{D} \quad (8.16)$$

Where, $\mathbb{T}$ and $\mathbb{D}$ are the allowed output and input sets. From (8.15), the optimization parameters can be reduced such that:

$$J_k(d_p, d_{p-1}, k, T_{p,n}) = \frac{1}{N_i} \sum_{j=k}^{k+N_i} \sum_{i=1}^M \lambda_i M(\theta_k^{(i)} \varphi_j^{(i)}) \quad (8.17)$$

Such that:

$$\hat{v}_k = \sum_{i=1}^M \lambda \hat{v}_k^{(i)} \text{ with } \sum_{i=1}^M \lambda_i = 1 \quad (8.18)$$

and ramp signal slope:

$$\Delta r_{k+1} = \begin{cases} d_p, & \frac{T_{d,p}}{n} \leq z \leq \frac{T_{h,p} + T_{d,p}}{n} \\ s_{APRBS} * z + d_{p-1}, & 1 \leq z \leq \frac{T_{d,p}}{n} \end{cases} \quad (8.19)$$

It is worth noting that the length of the input sequence can change during the signal's ramp period which is dependent of the signal's design points. Due to the adaptive nature of the signal, as well as the PV module's varying parameters and varying APRBS frequency, the dwell time, $T_{h,p}$ was considered as a design parameter such that, the optimal design points could be determined by the optimization problem as follows:

$$\Delta R_{APRBS}^* = \max_{\Delta R_{APRBS}} (\Delta R_{APRBS}, k, T_p) \quad (8.20)$$

With,

$$d_p^* = \max_{d_p} (d_p, d_{p-1}, k, T_{p,n}) \quad (8.21)$$

The criterion searches for the input sequence that causes the largest disagreement in the committee models within the APRBS signal's time horizon. This formulation can be illustrated as follows:

$$M_{QBC}(\beta_k^{(i)} \gamma_k^{(i)}) = \left\| f(\beta_k^{(i)} \gamma_k^{(i)}) - \sum_{i=1}^M \lambda_i f(\beta_k^{(i)} \gamma_k^{(i)}) \right\|_2^2 \quad (8.22)$$

Each optimal design point causes the largest variance of the model outputs. The steady state behavior of the APRBS signal was optimized by considering design points which leads to the minimal model output:

$$J_{opt}(d_p) = - \sum_{j=k}^{N_s} \left(\sum_{i=1}^M \lambda_i M(\theta_k^{(i)} \varphi_j^{(i)}) \right) \quad (8.23)$$

To optimize the transient state, an input ramp sequence which led to the lowest cost over the signal time horizon was determined by using (8.24).

$$J_k(\Delta R_{APRBS}, k, T_p) = - \sum_{j=1}^{N_s} \left(\sum_{i=k}^{k+N_i} \sum_{l=1}^M \lambda_l M \left(\theta_k^{(i)} \varphi_j^{(i)} \right) \right) \quad (8.24)$$

The APRBS signal was generated and shown in Figure 8.2 (a). Unlike the PRBS, the signal's amplitude is shown to be modulated around the fundamental frequency of the signal. The design points and signal dwell times are illustrated by Figure 8.2 (b). The figure shows a consistent slope, s_{APRBS} as defined by (8.19). The power spectrum APRBS signal in Figure 8.2 (c) shows a maximum usable bandwidth of 99 kHz. This shows how removing sudden changes of the PBRBS reduces the excitation of larger frequencies. Although this bandwidth is lower than the PRBS, it is still sufficient in tracking the PV module voltage.

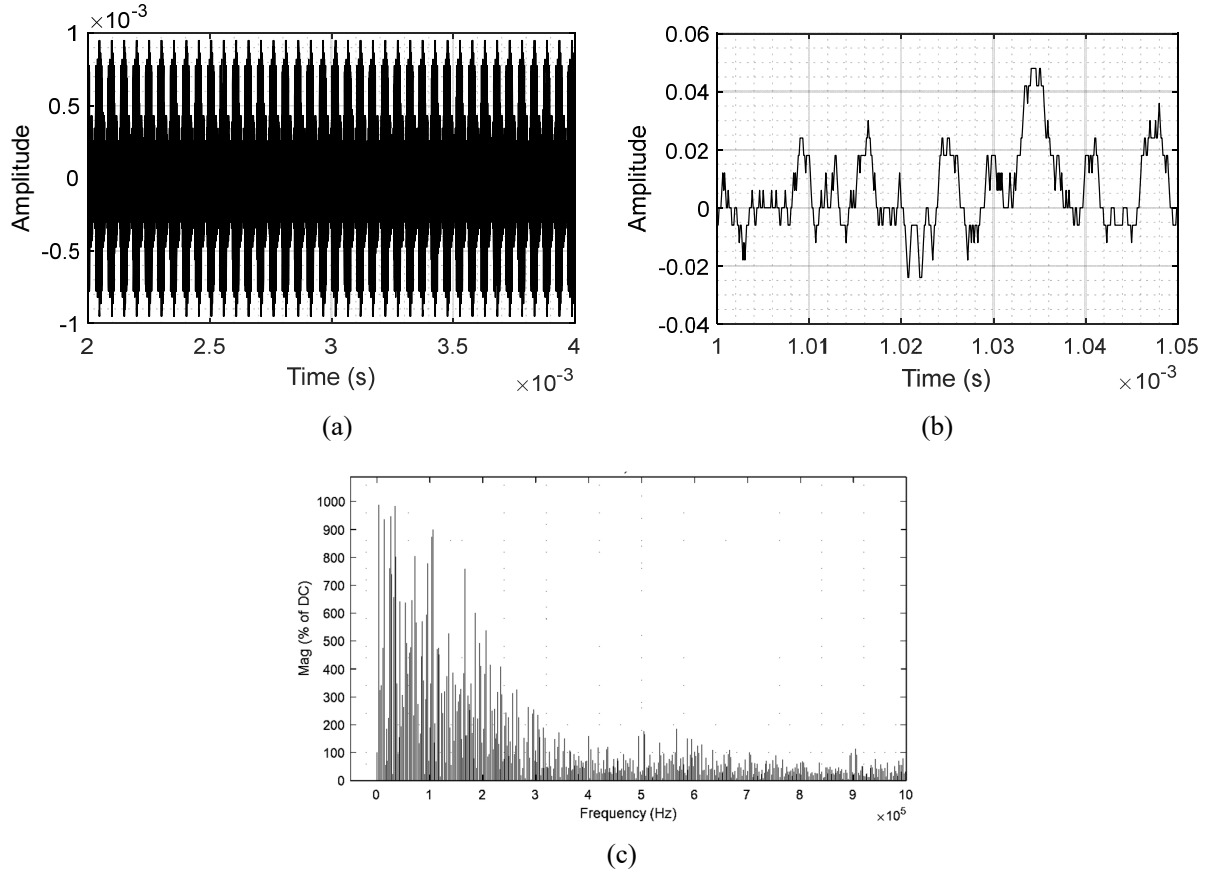


Figure 8.2 APRBS signal generation (a) 12-bit APRBS signal (b) APRBS slope (c) APRBS power spectrum

To evaluate the signal, each model committee consists of four multilayer neural networks in parallel configuration. The input and output are scaled such that:

$$u \in [0,1] \text{ and } y \in [0,1] \quad (8.25)$$

These are also delayed once to describe a first order difference equation. A good approximation of the system in high dynamic regions is necessary for high performance control – desired in the MPC, based voltage excitation signal design. The input space distribution for the signal design points,

illustrated in Figure 8.3, shows the interpretation of the design points input space in nonlinear autoregressive network with exogenous inputs (NARX) configuration.

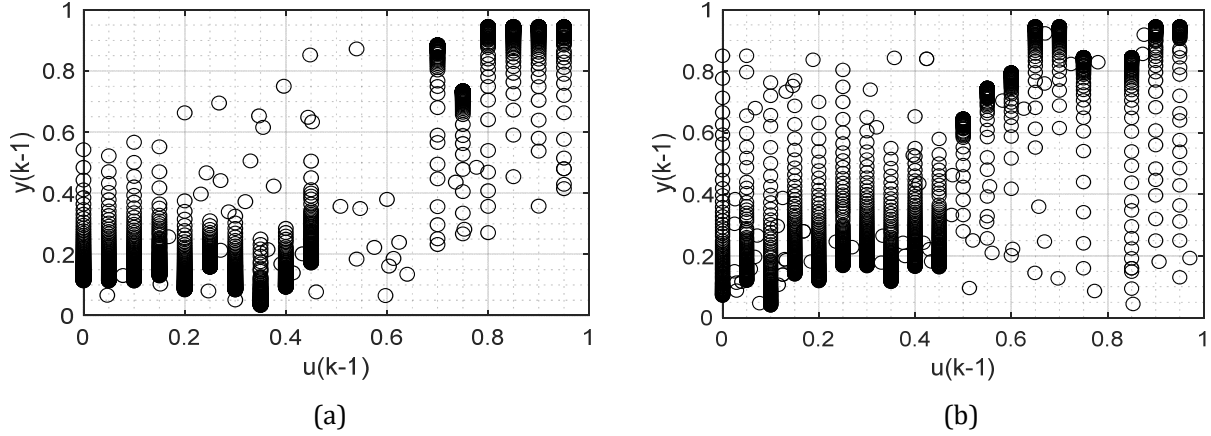


Figure 8.3 Signal input space distribution (a) PRBS (b) APRBS

From Figure 8.3 (a), the PRBS design points distribution show significant weakness in the signal's rate of constraint. The data points are distributed such that data is missing at the center and along the columns of the input space. The points are concentrated along the quasi-static excitation region – thus, resulting in less dynamic excitation for the PV system – particularly in variable conditions and frequency. Figure 8.3 (b) also shows a weak rate of constraint for the APRBS signal, as well as data points missing along the columns of the distribution space. However, although the design points are also concentrated along the quasi-static excitation region, the distribution shows a significant spread along the dynamic excitation region – thus significantly improving the signal's dynamic performance and its ability to keep the system within its linear region in varying frequency and variable point operation conditions.

8.4 Experimental results and discussions

8.4.1 Voltage excitation signal induction

To show how the modulated PRBS amplitude mitigates the non-linearity of the converter at changing frequencies, the excitation signals for both APRBS and PBRBS duty cycle modulating signals were simulated with the PV module operating at maximum power, and the results shown in Figure 8.4. Figure 8.4 (a) shows that the PRBS excitation signal has large amplitudes at low frequency regions, affecting the crest factor – more importantly, the signal is greater than the pseudo-linear, 9.45% of the DC value.

Figure 8.4 (b) shows that the APRBS perturbation can mitigate system non-linearities by inducing an excitation signal that has an average peak of 9.24% of the dc voltage, as designed for, and with a 1.25 crest factor across the excitation frequency bandwidth. This helps keep the system within its linear range for better impedance estimation.

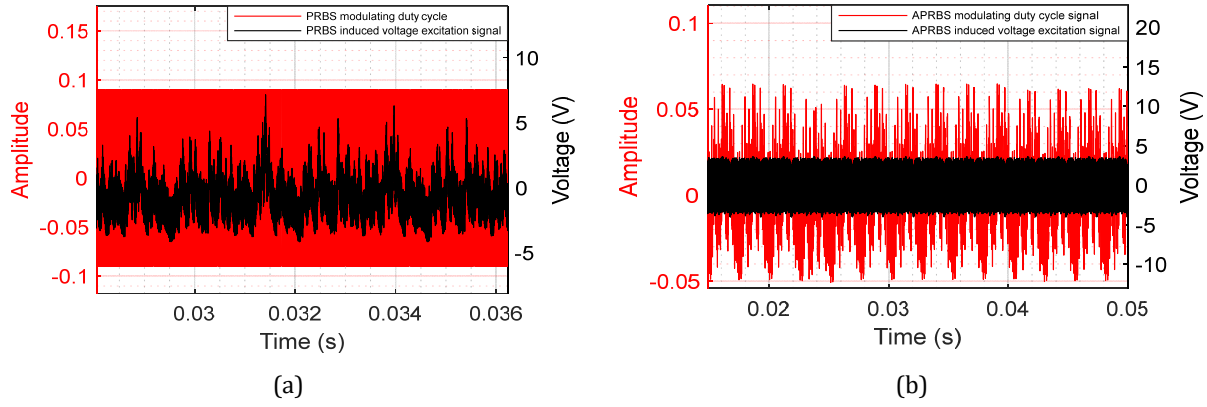


Figure 8.4 Experimental induced voltage excitation signals from modulating duty cycle perturbations (a) PRBS
(b) APRBS

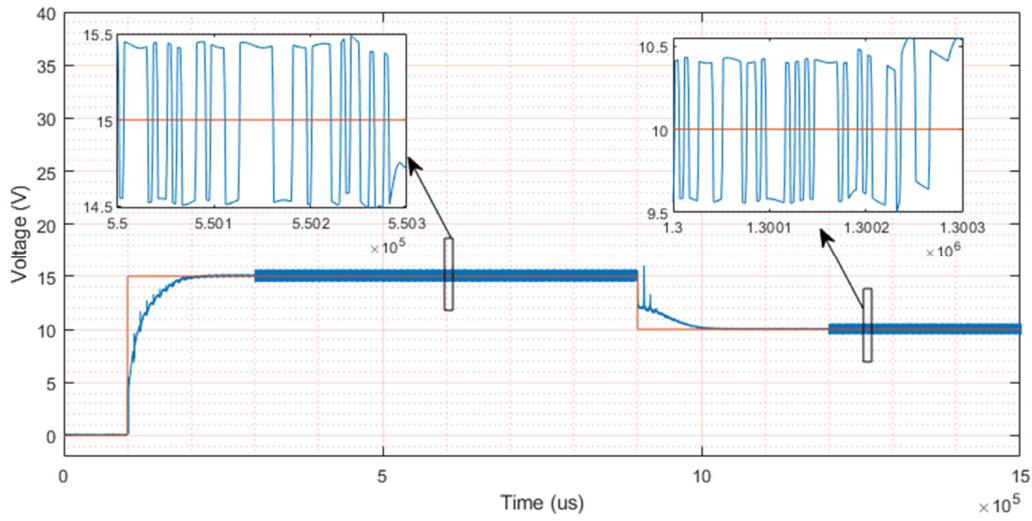
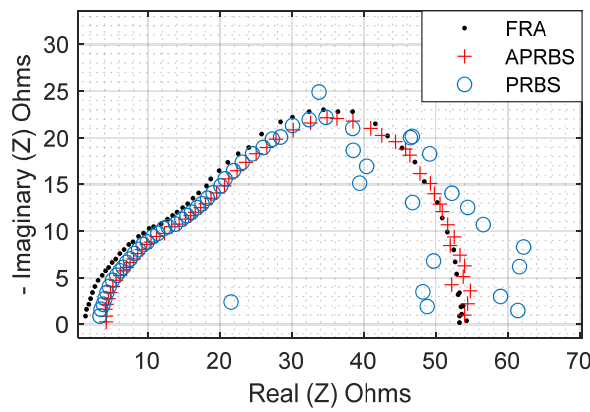


Figure 8.5 Induced APRBS voltage excitation signal resulting from APRBS perturbation, under input voltage variation

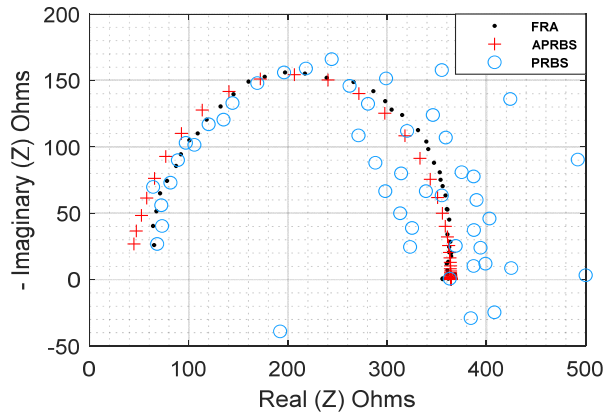
Figure 8.4 shows that the induced voltage excitation signal resulting from APRBS perturbation, has a better steady state response compared to Figure 5.20. The amplitude of higher frequency components has improved significantly, while the amplitude of lower frequencies has reduced; this resulted in a reduced crest factor of 1.47 – adequate for the purposes of the applications. The excitation amplitude for the lower voltage is now 10.1%, while its 8.3% for the 15V command – both within the desired range.

8.4.2 Impedance results

Figure 8.6 (a) and (b) show that for microcracks under high irradiance conditions, the APRBS performs adequately and with minimal error across all frequencies, when benchmarked against the FRA. The PRBS impedance estimation shows highly non-linear behaviour at lower frequencies because the excitation signal induced by the signal drives the PV module into its non-linear range of operation. The largest error between the FRA and APRBS exists at lower frequency parameters as shown by Table I – this is because the largest excitation amplitude error occurs at lower frequencies.

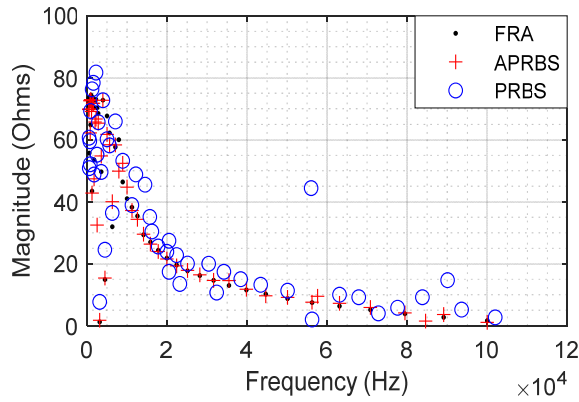


(a)

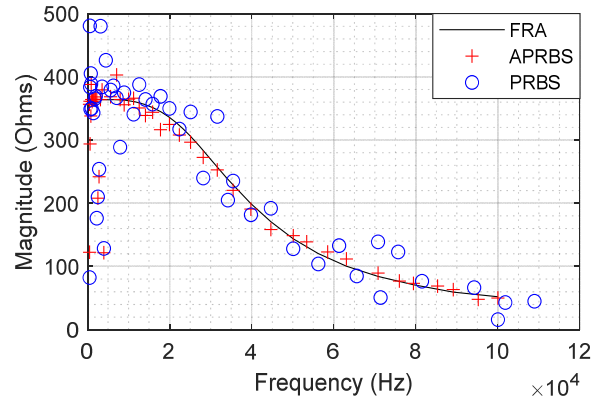


(b)

Figure 8.6 Nyquist APRBS impedance extraction of a PV module under microcracks and well-lit conditions (a) high irradiance (b) low light

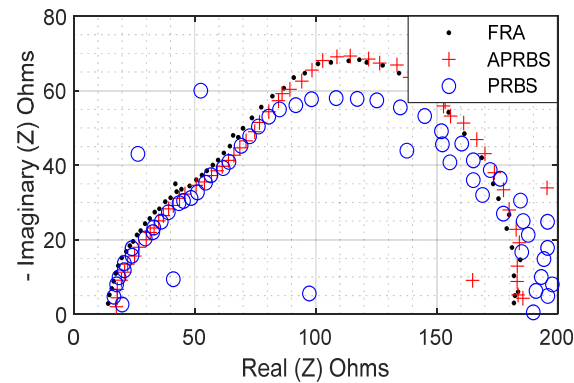


(a)

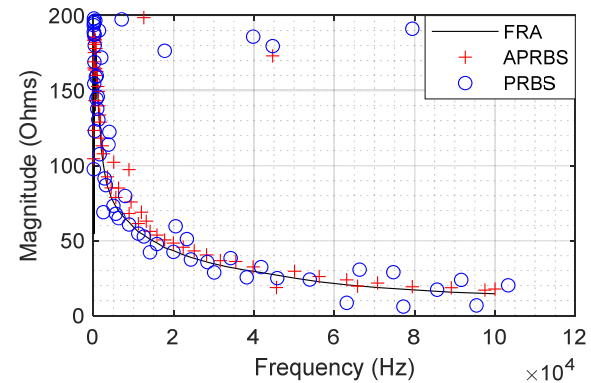


(b)

Figure 8.7 APRBS impedance magnitude of a PV module under microcracks and low-lit conditions (a) high irradiance (b) low irradiance



(a)



(b)

Figure 8.8 Impedance estimation during partial shading (a) Nyquist plot (b) Impedance magnitude

Figure 8.7 (a) and (b) show that the APRBS signal exhibits satisfactory results for microcracks under low irradiance conditions. Although the PRBS response shows better results, they are still inadequate for reasonable impedance estimation. It is worth noting that the response under poorly lit conditions is

first order, compared to the second order for high irradiance conditions. This can be attributed to active dynamics of the module cells while they are illuminated. The maximum error between the converter and the FRA is lower at low light conditions is illustrated by Table II. At low light conditions, the PV module's DC voltage and current are significantly smaller; this results in an excitation signal with a lower magnitude, which increases the APRBS signal's SNR and hence, increases the difficulty in signal detection. The partial shading impedance estimation, Figure 8.8 (a) and (b) show a second order response with a higher impedance than high irradiance conditions; this is because the impedance was estimated at a module voltage beyond the global maximum power point to show the behaviour of the impedance during this condition; this impedance error is shown in Table III.

Table 8.1 Error estimation under microcracks in high irradiance conditions

Parameter	PRBS maximum percentage error	APRBS maximum percentage error
r_s	0.06	0.69
r_{np}	15.23	1.30
c_{np}	10.23	2.01
r_{pp+}	1.01	0.96
c_{pp+}	1.24	0.74

Table 8.2 Error estimation under microcracks in low irradiance conditions

Parameter	PRBS maximum percentage error	APRBS maximum percentage error
r_s	0.11	0.27
r_{np}	16.36	1.12
c_{np}	9.25	1.36
r_{pp+}	—	—
c_{pp+}	—	—

Table 8.3 Error estimation under partial shading

Parameter	PRBS maximum percentage error	APRBS maximum percentage error
r_s	0.68	1.21
r_{np}	10.23	2.43
c_{np}	19.25	3.36
r_{pp+}	1.23	0.98
c_{pp+}	3.36	0.34

8.5 Chapter conclusions

In this chapter, an APRBS approach to converter-based BIS impedance extraction of PV systems was presented. The signal was designed such that it could mitigate frequency-induced non-linearities observed in the PRBS impedance because of induced excitation that comes with using a power converter as an excitation source. This was achieved by modulating the design points of the PRBS signal such that the signal maintains lower amplitudes at lower frequency values. Although the PRBS showed better impedance errors at higher frequencies, the APRBS signal showed good tolerance low frequency non-linearities and resulted in overall minimal impedance estimation. This helped ensure that the impedance changes were only due to changes in the PV module, and not design irregularities.

The results showed that at lower lighting conditions, the higher frequency parameters are difficult to achieve because of the larger impedance. It is important to note that this excitation signal echoes similarities to the parametric excitation signal discussed in the previous section; the difference being the digital configuration and implementation of the APRBS. However, both excitation signals are designed to respond aggressively to nonlinear behavior from the induction of the excitation signals through a nonlinear, dynamically responsive switch-mode converter. The last aspect to note is that due to the use of a power converter, the technique is more suited for diagnostics are good lighting to ensure the converter meets its minimum operational power requirements – where the converter’s rated power dictates the module that can be coupled to the system. The other major limitation is the cost associated with processing impedance information at high frequencies in real-time. Although this is significantly lower when using the APRBS sequence as opposed to the use of analog broadband signals.

8.6 References

- [1] S. Nejad and D. T. Gladwin, “Online Battery State of Power Prediction Using PRBS and Extended Kalman Filter,” *IEEE Transactions on Industrial Electronics*, vol. 67, no. 05, 2020.
- [2] M. T. Talat and A. H. Khawaja, “Li-Ion Battery Parameter Identification Using Pseudo Random Noise,” in *2019 15th International Conference on Emerging Technologies (ICET)*, 2019.
- [3] Jussi Sihvo, Tuomas Messo, Tomi Roinila, Roni Luhtala, “Online identification of internal impedance of Li-ion battery cell using ternary-sequence injection,” *Energy Conversion Congress and Exposition (ECCE)*, 2018.
- [4] Jussi Sihvo, Tuomas Messo, Tomi Roinila, Roni Luhtala, “Fast Approach for Battery Impedance Identification Using Pseudo-Random Sequence Signals,” *IEEE transactions on power electronics*, vol. 35, no. 03, p. 2548, 2020.
- [5] Shenton, K. Fang and A. T., “Constrained Optimal Test Signal Design for Improved Prediction Error,” *IEEE transactions on automation science and engineering*, vol. 11, no. 04, p. 1191, 2014.

- [6] R. Harley. Jung-Wook Park, "New External Neuro-Controller for Series Capacitive Reactance Compensator in a Power Network," IEEE transactions on power systems, vol. 19, no. 03, p. 1462, 2004.
- [7] M. Gringard and A. Kroll, "On the parametrization of APRBS and multisine test signals for the identification of nonlinear dynamic TS-models," 2016 IEEE Symposium Series on Computational Intelligence (SSCI), 2016, pp. 1-8, doi: 10.1109/SSCI.2016.7849984.
- [8] B. V. Alamdari, A. Fatehi and A. Khaki-Sedigh, "Neural network model-based predictive control for multivariable nonlinear systems," 2010 IEEE International Conference on Control Applications, 2010, pp. 920-925, doi: 10.1109/CCA.2010.5611265.
- [9] T. C. S. Wibowo, N. Saad and M. N. Karsiti, "The simulation of MISO MPC for gaseous pilot plant control with presence of measurement noise," 2007 International Conference on Intelligent and Advanced Systems, 2007, pp. 1107-1110, doi: 10.1109/ICIAS.2007.4658556.
- [10] M. Cui, X. Zan, F. Chi, M. Wang, and D. Yu, "Research on optimization of SRG power generation performance based on the QBC with coupled inductor," 2017 Chinese Automation Congress (CAC), 2017, pp. 3481-3486, doi: 10.1109/CAC.2017.8243383.
- [11] Z. He, K. Lv, S. Yu, D. Shi, and J. Yan, "QBC-Softmax Algorithm for E-nose Data Processing Based on Different Informativeness Evaluations," 2018 IEEE 3rd Advanced Information Technology, Electronic and Automation Control Conference (IAEAC), 2018, pp. 2085-2088, doi: 10.1109/IAEAC.2018.8577772.
- [12] Y. Zhao and Y. C. Cao, "An active TAN classifier based on vote entropy-maximum entropy of QBC," 2009 IEEE International Conference on Control and Automation, 2009, pp. 1587-1590, doi: 10.1109/ICCA.2009.5410440.
- [13] M. Bujarbaruah, U. Rosolia, Y. R. Stürz and F. Borrelli, "A Simple Robust MPC for Linear Systems with Parametric and Additive Uncertainty," 2021 American Control Conference (ACC), 2021, pp. 2108-2113, doi: 10.23919/ACC50511.2021.9482957.
- [14] S. Osawa, T. Nakano, S. Matsumoto, N. Katayama, Y. Saka, and H. Sato, "Fault diagnosis of photovoltaic modules using AC impedance spectroscopy," 2016 IEEE International Conference on Renewable Energy Research and Applications (ICRERA), 2016, pp. 210-215, doi: 10.1109/ICRERA.2016.7884539.

Chapter 9

Solar module impedance estimation through SMC-based inductor current ripple excitation

9.1 Introduction

As discussed in the previous chapters, the major drawback to using converters for systems with a wide bandwidth, like solar modules, is that for adequate measurements, the switching frequency of the converter must be about 5 times the maximum injected excitation frequency. As discussed in the introduction solar modules have high impedance bandwidths; this then requires high switching frequencies from the converter – which increases the cost and complexity of the system. Wide bandgap semiconductors, high sampling digital processing units, resonant gate drivers and fast recovery diodes are required to meet the high frequency standards of performing impedance measurements on the modules – which significantly drives up the cost of the system.

This chapter thus, investigates the use of a switch-mode converter's inductor current ripple as the excitation signal and its impact on the power regulation; with this setup, a traditional low cost, compact converter would be used for high frequency device excitation, without requiring expensive processing units or wide-bandgap semiconductors. In Chapter 4 and 5, results in both simulation and experimental results showed that it was possible to use the inductor current ripple to excite the module and get a good enough response signal. However, during the experimental results, it was shown that if the inductor current ripple is not designed adequately, it exceeds the desired maximum amplitude for both power regulation and device excitation, depending on the switching speed of the converter. It is worth noting

that for the purposes of this chapter, the injection of an externally generated perturbation duty ratio to induce the desired excitation signal, will be referred to as, external excitation.

9.2 Excitation system overview

To investigate the use of the inductor current ripple for characterization, a boost type DC-DC power optimizer was designed to implement EIS on a Shell monocrystalline solar module – this setup is shown in Figure 9.1. The multiphase configuration would not work because it significantly reduces the inductor current ripple – a parameter required for this solution.

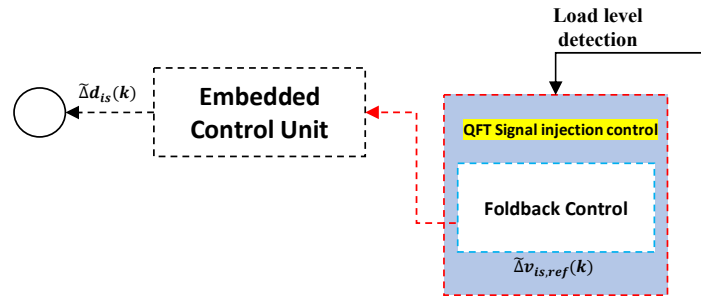


Figure 9.1 Inductor current ripple excitation system setup

To excite the PV module using the inductor current ripple of the converter, the output voltage $V_{pv}(t)$ and output current, $I_{pv}(t)$ are modelled as:

$$V_{pv}(t) = V_{dc}(t) + \Delta V_{sw}(t) \quad (9.1)$$

While the current is:

$$I_{pv}(t) = I_{dc}(t) + \Delta I_{sw}(t) \quad (9.2)$$

Where, $V_{dc}(t)$ is the module's DC output voltage; $I_{dc}(t)$ is the module's DC output current; $V_{sw}(t)$ and $I_{sw}(t)$ are the high frequency switching components due to the converter – now to be designed as the module's excitation signals. The inductor current ripple of the boost converter operating in CCM can be best described as a triangular waveform, whose symmetry is dependent of the converter's operating duty ratio. However, because of the symmetry, duty ratios above and below 50% are mirror images of one another, and the converter is designed to significantly boost low voltage, the description of the signal will focus on ratios 50% and above. Foldback control is implemented in conjunction with the MPPT control. It is implemented to allow for the variation of the switching frequency for impedance estimation while maintaining good efficiency – it does this regulating the charging current to the battery. This type of control is implemented digitally by measuring the load current and then varying the switching frequency of the converter based on the load level.

9.3 Boundary between external and current ripple excitation

Chapters 6 – 8 analysed characterization of the PV solar module through the induction of a voltage excitation from an externally sourced modulating perturbation duty ratio. The main takeaway from solutions of that nature was the requirement of larger switching speeds from the converter to achieve good resolution in the measurements at higher frequencies. As discussed in the introduction, the solution presented in this chapter is presented to alleviate these drawbacks. However, this solution isn't without its own drawbacks, the major being the characterization at low frequencies, i.e., problems that occur in the transfer of power at low switching speeds in the converter. The section thus, assess the impact of the converter's switching frequency on power regulation and impedance measurements.

9.3.1 *Choice of converter switching speed*

In choosing the most optimal switching speed for a switch-mode converter, a good design generally considers frequencies where the converter's static losses are approximately equal to its dynamic losses. This requires an engineering compromise from the designer to ensure that while performing impedance measurements, the dynamic behaviour of the system presents minimal deviations in the system's static behaviour. Switch-mode converters operate at high frequencies for several reasons; one of the major factors is the size of the components, more specifically, the inductor and capacitor sizes. At low frequencies the component sizes increase, as well as the ripple associated with both the inductor and capacitor. This significantly affects the amount of energy that can be transferred to the load cyclically per second. The acoustic noise of the boost converter also goes up at lower switching frequencies, particularly when the speed enters the audio range – that is, frequencies between 20Hz and 20kHz [1]. Most power converters are designed to operate at the boundary between the audible and inaudible range, that is, 20kHz ; this frequency point is good compromise because it is low enough such that switching losses can be minimised, and high enough such that the passive component sizes of the converter can be minimized as well.

Electromagnetic interference (EMI) at high switching frequencies is another factor; this is the production of electrical noise by the switch-mode converter that affects both the operation of the converter and electrical systems in close proximity [2]. Where the power transferred by the converter flows from source to load, the EMI noise flows from the converter to the load and back to the source [3]. This occurs because switch-mode converters employ passive components to isolate or transfer power between input and output at much lower frequencies than the switching frequency. Efficiency is another major in choosing the switching speed; although increasing the frequency may solve some issues at lower frequencies; at high frequencies, however, switching losses and hence, dynamic losses increase – this is particularly pertinent at low load current conditions. When the converter moves from full to light load conditions, it compensates for this by drawing more current from the supply; this increases the drain to source current and hence, losses across the switch.

To summarize, higher switching speeds in switch-mode converters are preferred over lower speeds. This is because the issues affecting the system at slow speeds outweigh those occurring at faster speeds. Having gone through the considerations, the various scenarios for the choosing either external or current ripple excitation, were considered. For the purposes of performing linear impedance estimation on the monocrystalline module, it appears around the peak impedance amplitude on a Nyquist curve, under high irradiance conditions. On the other hand, based on the results, the impedance at 100kHz was consistently located at the point of definition for module's series resistance; as such, as per the analysis in the previous section, the converter would thus need to be driven such that it operates at this switching speed.

9.3.2 Characterization under high irradiance conditions

In section 5.5.7, tests from various operational conditions were performed offline using the FRA. Focusing specifically on results under good lighting conditions i.e., irradiances $\geq 600\text{W}/\text{m}^2$ and peak imaginary ohmic response at excitation frequencies $> 20\text{kHz}$, it can be observed from Figure 9. and results from section 5.5.7, that the frequency distribution in the module's impedance is skewed towards higher frequency components i.e., the spread of higher frequencies almost fully characterizes the module. In fact, as observed from the graph, the highest imaginary component of the impedance on a Nyquist plot for the module occurs at frequencies well outside the audible range.

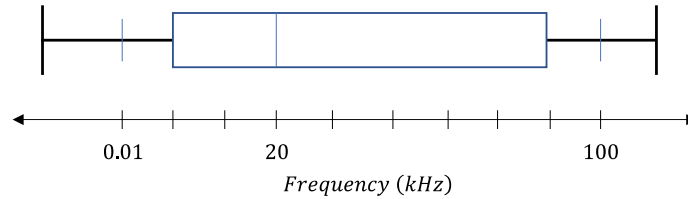


Figure 9.2 1D frequency distribution of the module's Nyquist response under high irradiance conditions

As discussed in the previous section, the most desirable switching speeds for the converter are greater than 20kHz . This means that under these conditions, the inductor current ripple can be employed to characterize at least half of the module's Nyquist response, but not the full spectrum. This is because the right half of the response occurs at frequencies lower than 20kHz – frequencies well below the speeds the converter will be designed to operate under. Thus, performing the full spectrum characterization of impedance measurements under high irradiance conditions requires both current ripple and external excitation because of the limitations in the switching speeds of the converter. It was established in earlier chapters that, when performing impedance measurements using an externally induced excitation signal, as per the traditional technique, the switching speed of the converter has to be at least 5 times the bandwidth of the device; this was in fact the primary inspiration behind the adoption of wide bandgap semiconductors – to improve the switching utilisation while performing impedance measurements from tradition silicon switches. With the effective bandwidth now reduced below 20kHz , means

the cost of the system can be drastically reduced, while operating at a good switching utilisation during impedance measurements.

9.3.3 Characterization under low irradiance conditions

Looking back at the results in Chapter 5.5.7, but now focusing on low irradiance conditions – irradiances between 100 W/m^2 and 600 W/m^2 . The frequency distribution under these conditions ranges from achieving symmetry in the number of frequency points attainable on the left and right side of the maximum imaginary ohmic response, to skewing towards lower frequencies; this is illustrated in Figure 9.3.

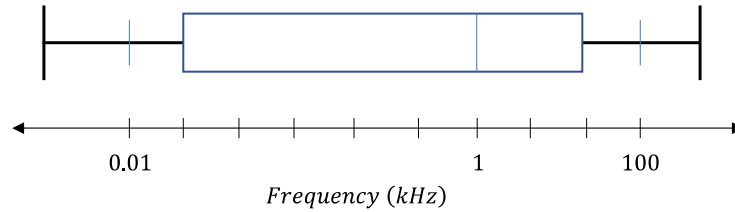


Figure 9.3 Frequency distribution of the module's Nyquist response under low irradiance conditions

The figure shows that the frequency distribution has now shifted to the left of the curve, favouring lower frequency components in the excitation. The peak imaginary ohmic response is now located at frequencies below 5 kHz . Evidently, under such conditions, inductor current ripple excitation would only result in the characterization of less than 50% of the module's Nyquist response. As observed from Figure 9.4, these conditions typically occur before 8:00am in South Africa [4]; during this time, most solar systems are not considered to be in their most optimal solar window. Solar plants are typically designed achieve maximum performance within a solar window between 9:00am until 15:00pm [5], as such, characterization under these conditions must be designed with these factors in mind. In conclusion, external excitation is more suited for these conditions more than current ripple excitation. In applications where the impedance information at the higher frequency end is more pertinent, then this solution would be the most optimal because as seen from results in section 5.6.7, high irradiance conditions contain more high frequency points than lower frequency points.

Anything below 100 W/m^2 is considered dark and it is common practice in solar power plants is to characterize PV modules under dark conditions; this is because these conditions are able to provide extra information about the module, while having considerably less noise from the light intensity and electron mobility [6]. Dark I/V measurements are performed by injecting carriers into the system through power electronics, as opposed to making use light generated carriers [7][8]. The power optimizer is designed such that under dark conditions, when the batteries have a large potential than the modules, the batteries do not discharge through the modules. As ideal as dark conditions are for characterization, they cannot be performed through online characterization, nor can they be performed using a switch-mode converter – the power from a non-illuminated module would not be enough to

operate a power converter. Thus, the solution provided in this chapter would not be applicable to dark conditions.

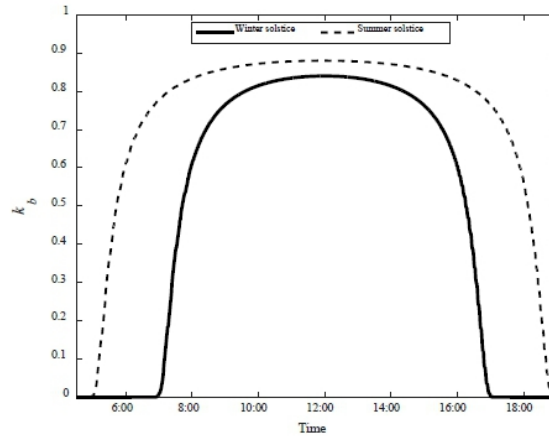


Figure 9.4 Durban direct irradiance profile during the winter and summer solstices [5]

9.4 Design considerations for the current ripple excitation magnitude

As with any excitation signal, careful attention should be paid into its design to ensure optimal device excitation. This section assesses the inductor current ripple of the converter in terms of its definition in the context of power converter to optimise its excitation magnitude; and then finally, its frequency response to better understand its relationship between the switching frequency and its magnitude. In CCM, the converter's current ripple is as described by, (4.3); ideally, this peak-peak inductor current ripple magnitude is dependent on several power stage parameters, the converter's operating point, the module's operating voltage, the size of the inductor used, and the system's operating frequency. To get good impedance values, the power stage parameters need to be chosen such that they maximize (4.3). Under ideal conditions, increasing the time-averaged duty ratio, will decrease the term, $(1 - D_{dc})$ thus increase the peak-peak magnitude of the current ripple. Increasing the modulating ratio thus, not only ensures that the converter remains in CICM but, increases the ripple's magnitude as well. The best magnitudes are obtained at larger operating voltage values for the module; again, this helps keeps the module operating in CICM. Small inductance values yield the largest ripple magnitude. However, there is a limit to how small the inductor can be because the amount of energy the inductor can store depends on the size of the inductor. In other words, as the primary storage element, the inductor must be large enough to allow for seamless power transfer between the input and output. Lastly, the frequency is required to be as low as possible to achieve good ripple magnitudes. However, to achieve optimal impedance estimation, the perturbation frequency needs to be, at least as large as the module's bandwidth. This creates a conflict that will be discussed in more detail in the next section.

$$|\Delta I_{sw}\{D_{mppt}\}| = \frac{\overline{V_{pv}}}{L_a f_s} \left[\frac{(D_{dc}(1 - D_{dc}))^{-1}}{\left(\frac{r_{La}}{|\tilde{\Delta}Z_{bat}|(1 - D_{dc})^2} \right) + 1} \right] \quad \forall D_{mppt} \mid D_{mppt} \in [0,1] \quad (9.3)$$

Equation (9.3), however shows that when considering an ideal inductor, its DC parasitic resistance (DCR) affects the magnitude of the current ripple. The effects of this are illustrated in the figure as follows:

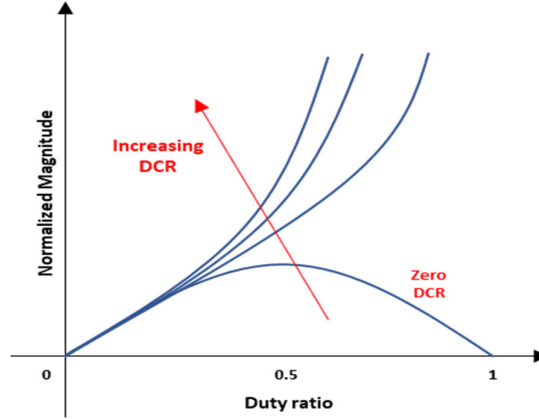


Figure 9.5 Effects of the inductor's parasitic resistance on the magnitude of the peak-peak inductor current ripple

Compare to inductor current ripple obtained in Figure 4.8, Figure 9.5 shows a more practical variance of the inductor ripple as the modulating duty ratio changes. Figure 9.5 shows that when the DCR is zero, the results in Figure 4.8 are obtained; however, as the DCR increases while the load remains the same, the response increases to infinity as the duty ratio approaches unity – getting steeper as the DCR to load ratio increases. This means that to avoid inductor current ripple values that bring the converter close discontinuous operation, the DCR has to be small compared to the lightest anticipated load value.

A good design generally considers a good compromise between the inductor and capacitor sizes; this is achieved by choosing an inductor current ripple such that $\Delta I_{sw} \in [10\%, 40\%]I_L$ of the maximum input current. Moreover, the inductor size is designed such that the input current is continuous for load currents above 5% – 15% of full load current. This ripple flows in the output capacitor because of its low high frequency impedance; however, the ESR of the capacitor will produce a ripple that will flow through the load. Therefore, the output capacitor needs to have a high ripple rating in low frequency applications. The minimum current required for continuous mode also needs to be considered since impedance measurements must be taken in continuous mode to meet LTI constraints. It is worth noting that a higher ripple current increases the inductor proximity and core losses. However, a higher value of the inductor, with a low ripple, increases the DC resistance losses.

9.5 Design considerations for the excitation frequency

As mentioned earlier, careful attention should be paid to the switching frequency of the converter as it plays the most crucial role in designed the most optimal excitation signal. The measurement frequency range was chosen such that, when the frequency is varied and the peak-peak inductor current ripple changes, it meets both the linear amplitude range, ΔA_{is} and the optimal inductor current ripple range, ΔI_{sw} ; this means an intersection between the sets $\Delta A_{is} \cap \Delta I_{sw}$. This design allows for the switching frequency to be varied while providing minimal power regulation interference. With an inductance of $26\mu H$; output voltage of $38V$; and an input voltage of $20V$ chosen, the frequency can be varied while the inductor current ripple is observed. The module used has an $18V_{mp}$; for EIS measurements, the ECU pulls that voltage up to $20V$. While ideally, this voltage is to be kept constant during the measurements, a varying inductor current ripple will cause the input voltage to vary slightly. The control system then ensures that the output voltage remains regulated during the measurements.

In section 5.6.5 it was obtained that ΔA_{is} for the module is $8.51\% - 13.45\%$ of the module's DC voltage. Thus using (9.3) and the specifications mentioned above for ΔI_{sw} , the intersection of the sets, $\Delta A_{is} \cap \Delta I_{sw}$, yields a measurement set of $\Delta I_{eis} \in [10.51\%, 33.45\%I_L]$ for the current ripple. With an operating current of $3A$ at $20V$, the ripple translates to, $\Delta I_{sw,eis} \in [315, 1004]mA$. The response of the module shows how the small signal voltage responds to the inductor current ripple as the EIS excitation signal. This is an important consideration at higher frequencies where the response is attenuated significantly, for 3 reasons: firstly, the smallest amplitude of the response must have a good SNR; secondly, the smallest amplitude must be sufficiently large to overcome the limitations of signal processing equipment and data acquisition; and lastly, ensuring that lower frequencies aren't amplified such that the operating voltage decreases to, V_{mp} and current increases to, I_{mp} – as this will lead nonlinear EIS operation. With this variation, the EIS voltage response, is in the range $\Delta V_{sw,eis} \in [0.5, 1.5]V$. This shows that during the measurements, the input voltage can vary such that: $\Delta V_{dc,eis} \in [18.5, 20.5]V$. These voltage ranges are within the minimum and maximum input voltages for the boost converter designed, but most importantly keeps the PV operating as a voltage source during measurements. From (3), the measurement frequency range was then determined by varying the frequency such that the ripple remained within the measurement set, $\Delta I_{sw,eis}$ – this was a frequency range of $\Delta f_{eis} \in [21.47, 103.14]kHz$. This range represents the switching frequencies that are within the design limitations of the components chosen for the design. As mentioned earlier, continuously operating high frequencies under lights loads affects the converter's efficiency, to solve this many DC-DC incorporate frequency foldback mode. In this mode, the control system is designed to proportionally reduce the system's switching frequency as the load gets lighter. A similar approach will be employed when varying the switching frequency for impedance measurement purposes.

From equation (9.3), it can be observed that different input and output voltages will lead to a new set of measurement frequencies. Choosing higher input or output voltage increases both the allowable maximum and minimum measurement frequencies, while choosing lower values of either, does the opposite. It is worth noting the range of the set changes as well. Choosing a lower inductance value significantly increases the set range, as well as the maximum frequency, and vice versa. To increase the maximum frequency while decreasing the minimum frequency, higher values for either the input or output voltages and the inductance must be chosen.

9.6 Performance assessment during foldback control

Earlier in the chapter, the design considerations around the choice of converter switching speed were discussed. The performance assessment for a system designed to switch at $20kHz$, now operating at $100kHz$ was analysed by looking at the change in the efficiency, input current ripple, output voltage ripple, transient response, and loop stability in closed loop control.

9.6.1 System efficiency

To assess the efficiency of the system with increasing frequency, the losses associated with frequency changes are characterized i.e., switching losses. These losses, as define by (9.4), occur during the transition of the semiconductor between its ON-state and OFF-states. If a linear transition between the 2 states is assumed, then the losses are a linear calculation of the area of overlap between the transitions; however, in practical semiconductors, it is nonlinear, as illustrated in [9].

$$P_{sw} = \frac{I_{ds} f_{sw} \int_0^{T_{sw}} V_{ds}(t) dt}{T_{sw}} t_{on} \quad (9.4)$$

Where, I_{ds} is the drain source current; V_{ds} is the drain to source voltage; T_{sw} is the switching period of the converter; and t_{on} is the time it takes the drain to source voltage to return to zero. Based on this, the switching losses should increase by 64% as the converter increases its switching frequency. Conduction losses are another factor to consider. Although these losses are not directly frequency-dependent, they are dependent on the converter's duty ratio. These are losses that are due to the presence of the semiconductor's ON-resistance. At lighter loads, the switching speed is decreased in the converter; however, a lighter load increases the operating duty ratio, which in turn increases the conduction losses, hence, the switching speed can indirectly affect the conduction losses.

9.6.2 Input current ripple

To assess the theoretical change in the ripple as the frequency range changes from $20kHz$ to $100kHz$, equation (4.3) was considered. Under constant loading, the change in the current ripple should theoretically reduce by 80%. Assuming all other parameters in (4.3) remain constant, a converter designed at a switching frequency of $20kHz$ will have its inductor current ripple reduce by 80% when the switching

speed is increased to 100kHz , while keeping the system parameters constant. This is clearly well below the minimum percentage recommended for the inductor current ripple. At this percentage, the inductor would be greatly undersized for that ripple and the converter would have a significantly poor transient response to load step and release. If a ripple percentage of 40% is chosen for the design at 20kHz , the inductor current ripple would reduce to 8% of the module's rated current, should the switching speed increase from 20kHz to 100kHz . Clearly, choosing the higher end of the percentage is better to ensure good high frequency perturbation.

If the load changes from full load to lighter loads, the converter reduces the module's operating current. Since the inductor current ripple is the same, it brings the converter closer to discontinuous operation. If the converter were designed for 100kHz , this would be a major issue. Therefore, designing the converter for 20kHz operation is ideal to ensure continuous inductor operation as impedance measurements are taken. Thus, when performing impedance measurements, the control system will be designed such that it regulates the charging current to the battery load to ensure that a level of power efficiency is maintained while engaging impedance estimation. This will, again be discussed further considering the experimental results.

9.6.3 Output voltage ripple

The output voltage ripple was discussed in detail in Chapter 4. It is inherent to a switch-mode converter because of its switching states – more specifically the inductor current ripple. This ripple is designed to flow exclusively through the output filter capacitor because the time-varying current cause a perturbation of the capacitor potential. The capacitor ESR plays an important role in the output voltage ripple that is pertinent to the practical application of the converter; equation (9.5) illustrates this more clearly [10].

$$\Delta V_{bat} = \Delta I_{sw} \left(\frac{1}{8C_a f_{sw}} + \frac{r_{ca}^2 C_a f_{sw}}{2D_{mppt}(1 - D_{mppt})} \right) \quad (9.5)$$

As the frequency increases, the output voltage ripple decreases significantly. However, this is true without factoring in the effects of the ESR and load conditions. Under light load conditions, the operating speed of the converter is reduced for better efficiency – this will thus increase the output voltage ripple. To better manage this issue, the converter will employ pulse skipping mode; this will ensure that the value of the ripple is dependent on the sizes of the inductor and capacitor, with the frequency having little effect on the ripple. Under heavier loads, the converter employs PWM and operates at higher speeds and the output voltage ripple is thus, a function of the voltage spike due to the capacitor's ESR and the inductor current ripple.

9.6.4 Load transient response

The transient response of the load is another key performance factor that measures how long it takes the system to respond to load changing commands. The load step size is a major factor in the response, the larger the step size the longer it takes the system to respond. The other key factors are the sizes of the inductor and capacitor – more specifically their relation to the system's LC response. Apart from designing the both the inductor and capacitor to transfer adequate energy from the input to the output, they should be designed such that the response of the LC combination meets the transient requirements by the load.

The switching speed of the converter does play an even bigger role in managing this response. When the system is required to engage a heavy load i.e., during load step-up, the switching frequency and duty ratio increase to reduce the transient time it takes the inductor current to meet the load demand. Therefore, at high frequency the system has a better transient response under heavy loads. Under light loads the converter reduces the switching frequency; however, there exists a delay, which is worst at t_{on} ; better illustrated as, D/f_{sw} , from load release due to the onset of the timer of the on-time. This illustrates better transients at higher switching frequencies.

9.6.5 Loop stability

The loop stability of a boost converter is assessed through the characterization of its gain and phase margins. The gain margin is the gain which can be adjusted until the system becomes marginally stable; crossing this threshold then leads to system instability. It occurs at the phase crossover frequency – the phase at which the response reaches 180° of phase. The phase margin, on the hand, is the amount of phase the system can gain or lose before the system becomes marginally stable. It occurs at the system's gain crossover frequency – the frequency at which the system gain is unity. For good stable system, both gain and phase margins should be positive such that, the gain crossover frequency is lower than the phase crossover frequency. From this information, the stability of the system when it executes foldback control during impedance measurements was assessed.

For a system designed to achieve a particular phase margin at the gain crossover, the rule of thumb is that the maximum gain crossover frequency occurs between, 10% – 20% of the converter's switching speed; this is because of the limitation in the system's phase and transport lag. However, as discussed during the control design in Chapter 4, a boost converter experiences undershoot in its output voltage while operating at high duty ratios when the load demand occurs at higher frequency than the inductor transients can handle – transients attributed to a zero in the right half plane. To curb this issue, the boost converter's response is intentionally slowed such that the inductor has sufficient time to build up its energy. Thus, to ensure a good phase margin, the general rule of thumb is that the maximum crossover frequency should occur at frequencies less than 30% of the frequency at which the right half plane is located. To ensure that converter can overcome oscillation at the converter's resonance frequency, it

must have sufficient gain at this frequency i.e., the resonant frequency should be less than at least 30% the gain crossover frequency. With this mind, for a system designed for a lower frequency than the maximum impedance measurement frequency will remain stable during the process, however the reverse scenario would lead to loop instability; therefore, the converter must be designed for a lower frequency than the maximum perturbation frequency required for impedance estimation.

9.7 Optimal frequency points

Having determined the optimal frequency range for both optimal power regulation and impedance estimation, the optimal frequency points were now considered. These frequency points are chosen such that each of the impedance parameters can be estimated with the minimal number of points. In so doing, the total cost of the system of the system could be significantly reduced. These considerations are discussed in the context of each of the module's impedance parameters.

9.7.1 Series resistance

The real ohmic series resistance definition and its impact were discussed in detail within various subchapters during Chapter 2. Based on the FRA results obtained in chapter 5.5.7, for the monocrystal-line module under test, a response predictor variable in linear regression model can be linearly predicted with high accuracy only at high frequencies, $\cong 100kHz$, thus decreasing the uncertainty associated with the module's series ohmic resistance at these frequencies. Definitively, a single frequency point is required, at a minimum, to fully characterize the series resistance of the module. This means that the converter switching at 1 frequency and operating, as per its design, would be sufficient, without the need for foldback control.

To analyse how close this frequency is to the actual of the series resistance, consider an arbitrary high frequency point on a Nyquist plot, $P_H = \{r_{hr}, jr_{hi}\}$; where, r_{hr} is the real component, and r_{hi} the imaginary component at that frequency. By the definition of the series resistance on a Nyquist plot, the closer r_{hi} is to zero, the higher the accuracy and hence, a higher certainty that, $Re\{P_H\}$ is the module's series resistance; that is, $\forall |Im\{P_H\}| \cong 0, Re\{P_H\} \cong r_{is}$.

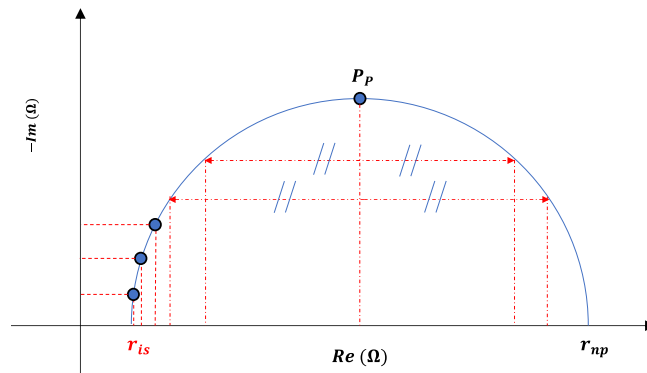


Figure 9.6 Graphical analysis of a typical Nyquist impedance representation for a first order electrochemical device

Figure 9.6 illustrates that in situations where, $|r_{hi}|$ has a sufficient magnitude, the likelihood of $Re\{P_H\}$ representing the device impedance is dictated by slope at which the curve approaches the real axis; that is, the steeper the slope of the curve, the closer $Re\{P_H\}$ is to r_{is} . However, this means that at least 2 frequency points are required to achieve high accuracy in the approximation – the point at the highest measurement frequency and another, closest to the highest frequency point. From these points, the slope of approach, $\angle P_s = \langle [r_{hi} - r_{hi+1}] / [r_{hr} - r_{hr+1}] \rangle$ can be determined. The intercept made by the slope with the real axis, yields an approximate value for the series resistance, r_{ss} . However, because of time constant in the capacitive nature of the response, $r_{hh} \geq r_{is} \geq r_{hr} \forall f_{is}$ within the module's bandwidth.

Evidently, $\lim_{\angle P_s \rightarrow \infty} r_{hh} = r_{is}$; this is better illustrated by the results in section 5.5.7. The steeper the slope of the curve, the closer, r_{hh} is to the series resistance value, r_{is} . Under conditions of low light to dark conditions, the slope of the curve approaches infinity and thus, r_{hh} is infinitesimally close to, r_{is} . Under high irradiance conditions, however, the slope of approach is shallower, meaning that there will be a higher uncertainty in the series resistance. Small values of this resistance are expected under these conditions, however; this means that the change in the actual vs the approximated value will be within a small range that may be considered negligible under certain applications. The most optimal solution to determine the series resistance is thus, using current ripple excitation rather than external excitation.

9.7.2 Diffusion resistance

This resistance is the difference between intersections of the curve with the real axis at the low and high frequency ends. There are 2 approaches that can be undertaken to determine the most optimal frequency points for this resistance. The first approach takes cues from its graphical definition; it requires a point at the highest frequency and one at the lowest frequency. The resistance would thus be: $r_{np} = Re\{P_L\} - Re\{P_H\}$; where, P_L is the impedance point at the lowest frequency. The highest frequency point, P_H was analysed in detail in the previous section. The lowest frequency point can be analysed in a similar manner to determine, r_{ll} the intersection of the curve with the real axis at the high frequency end. With this approach, both current ripple and external excitation would be needed to fully characterize the resistance – clearly not the most optimal solution.

The second approach requires using symmetry and the peak imaginary ohmic response. Again, this requires at least 2 frequency points: the peak imaginary response and either 1 high or low frequency point. To illustrate this, consider a frequency point at the peak of the response, $P_P = \{r_{pr}, jr_{pi}\}$. The diffusion resistance becomes: $r_{np} = 2|Re\{P_P\} - r_{is}|$, or $r_{np} = Re\{P_H\} + Re'\{P_H\}_{|Re\{P_P\}}$ i.e., the sum of the high frequency point's real component and its reflection about the real component of the peak imaginary ohmic response, and similarly using the low frequency point, $r_{np} = Re\{P_L\} + Re'\{P_L\}_{|Re\{P_P\}}$. The second approach clearly accommodates the use of either current ripple or external excitation for the resistance and is thus, considered optimal.

9.7.3 Diffusion capacitance

The diffusion capacitance is easiest to characterize. The impedance of a capacitor at the highest point in the curve is, $Re\{P_P\} = 1/(\omega_{is} c_{np})$; where, c_{np} is the diffusion capacitance in question. Thus, only a single frequency point is required for the characterisation of the capacitance. Under high irradiance conditions, current ripple excitation must be employed and under low irradiance conditions external excitation.

9.8 System design for inductor current ripple excitation

As mentioned in the introduction, with inductor current ripple excitation, a single-phase boost converter was design for the application, as such, the new specifications for the design are detailed in Table 9.1. The system power remains unchanged because it is the same module being tested, as well as the same load. The switching frequency, however, was reduced from $500kHz$ to a nominal $30kHz$ (which varies depending on the excitation required – as explained during the analysis) for 2 reasons; firstly, using the inductor current ripple means direct excitation i.e., the switching frequency is the same as the excitation frequency; hence, the maximum switching frequency required is $100kHz$; secondly, the nominal switching frequency was designed to be lower than the minimum switching frequency because, under changing load conditions, the foldback control algorithm needs to reduce the frequency to ensure both good efficiency and linear excitation.

Table 9.1 Inductor current ripple excitation system power regulation specifications

Specification type	Specification
System power	80W
Input voltage range	10 – 21V _{DC}
Nominal switching frequency	30kHz
Minimum switching frequency	20kHz
Maximum switching frequency	100kHz
Rated module current	5.35
Max DC load/charging current	2.68A
Inductor DCR r_L	100mΩ
Capacitor ESR r_C	30mΩ
Output voltage ripple	≤ 2% of V_o
Nominal inductor ripple Δi_L	25% of $i_{L,max}$
Maximum duty cycle	83.69%
Nominal duty cycle	50%
Minimum duty cycle	16.30%

If the nominal was chosen as the minimum, during foldback, the control would reduce the switching speed such that the linearity of the excitation magnitude is not satisfied. Using equation derived in Chapter 4, the inductor was designed to be $157\mu H$; output filter capacitor designed to be, $481\mu F$; the input filter capacitor, on the other hand, was chosen, based on to be $1.05mF$, closer to the original value because its dependence on the switching frequency is minimal compared to its biggest contributor, the module power – which remains the same.

9.9 Experimental results and discussions

Various results that illustrate the system excitation are discussed to assess the feasibility of the inductor current ripple as an excitation signal for a monocrystalline solar module.

9.9.1 Current ripple variation and foldback control results

The results illustrating the variation in the converter's excitation current magnitude with varying switching speeds were assessed in Figure 9.7. The load was kept constant during the test.

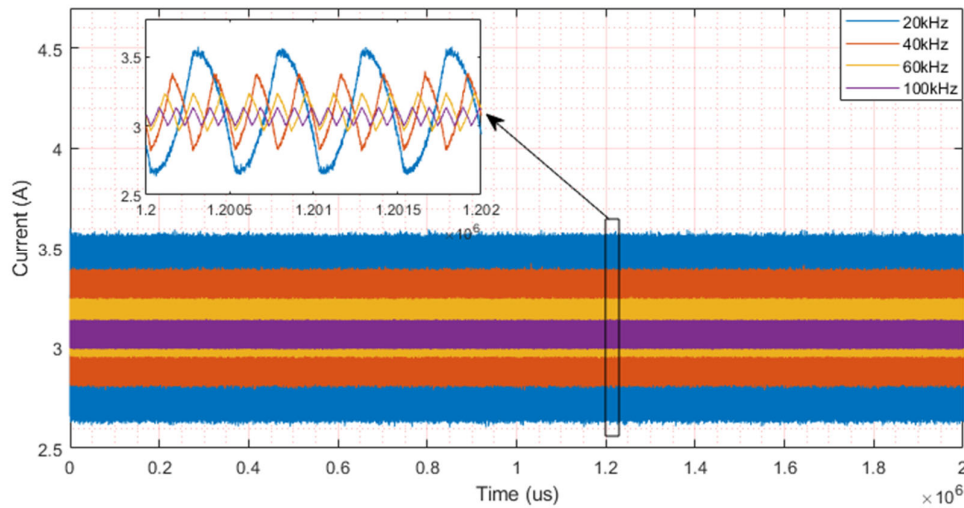


Figure 9.7 Boost converter input inductor current ripple at various switching speeds

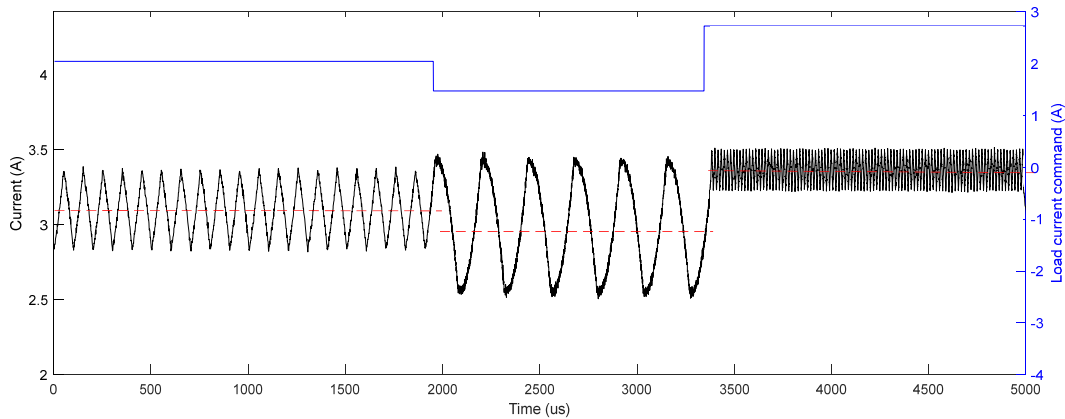


Figure 9.8 Foldback control of a boost converter's inductor current ripple

As illustrated by the figure, the inductor current ripple decreases with increasing frequency, as expected. At 20kHz , the peak-peak is 32% of the module's DC output current – within the desired specifications for both device excitation and power regulation. At 100kHz , however, the ripple is 9.73% of the DC current; this is lower than the desired 10.51%. The implications of this will result in a higher uncertainty in the characterization of the module's series resistance, as will be discussed. Foldback control was used to control the converter's switching frequency, under load changes, for a well-regulated inductor current ripple. The results of the control are discussed below. Three load commands were performed, light, nominal, and rated load. Figure 9.8 shows that under light loads, the control reduces the switching speed of the converter. The switching speed at the lightest load was 20.35kHz – with a magnitude that is 32% of the module's DC current, as discussed in the previous section, thus within both linear excitation and power regulation specifications. The control was designed to ensure that under lightest anticipated load, the control action does not fall below a modulating value that leads to switching speeds lower than 20kHz . Under nominal conditions, the switching speed control was 29.89kHz – as designed for – with a magnitude that is 27% of the module's DC current, and within the designed specifications. Under rated load, the switching speed was increased to 87.23kHz which helps regulate the voltage under heavy loads.

9.9.2 Impedance results

When performing impedance measurements using the current ripple as the excitation signal of choice, the frequency distribution of impedance response dictates the most optimal measurement solution. The behaviour of the electrons under different lighting conditions is well reflected in the results, and most importantly, in their mobility as impedance varies. value is, the higher the accuracy. However, at high frequencies the slope, $dIm/dRe \gg 1$, as such the real component at that condition will be approximate to the series resistance. To characterize the series resistance, the highest measurement frequency was used as the converter's switching frequency while both irradiance and temperature, and hence, input and output voltages, were varied. The accuracy of the series resistance was evaluated by observing the imaginary impedance value at each condition – this was compared to the series resistance obtained from the I/V curve. To characterize the module, the converter was operated at various switching speeds, exceeding 20kHz for reasons discussed earlier.

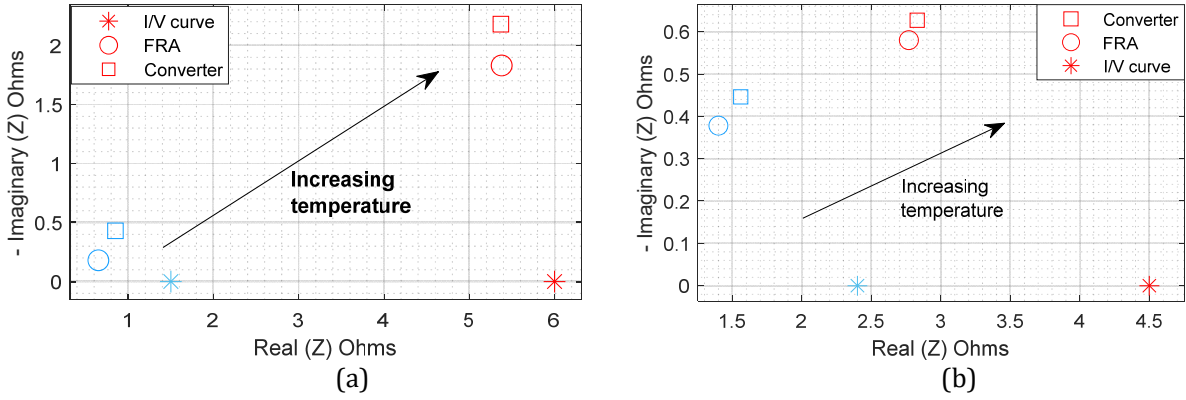


Figure 9.9 Series resistance under variations in (a) irradiation - well lit (red) low light (blue) (b) temperature - low (red) high (blue)

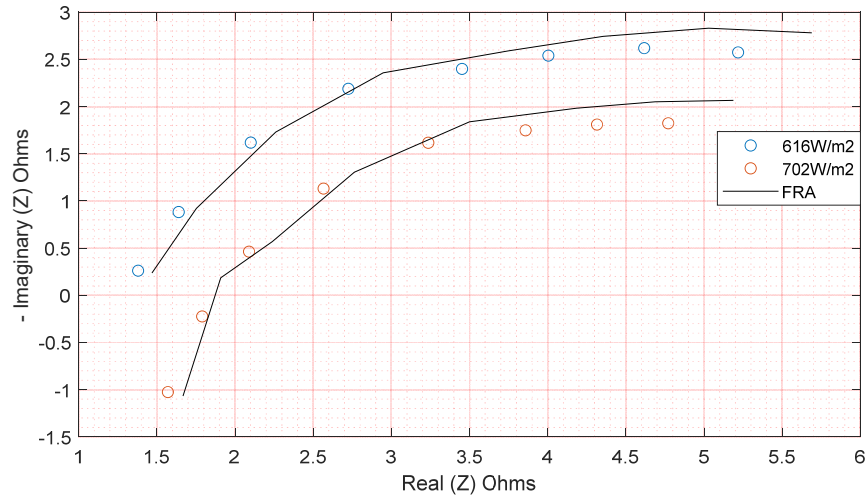


Figure 9.10 Inductor current ripple impedance estimation under high irradiance conditions

Figure 9.9 (a) shows that under high irradiance conditions, series resistance obtained from the I/V curve and EIS are within 5% of each other and within 7% from the FRA for the converter and 21% for the I/V curve. Under low light conditions, the results from EIS show a 6% error, while the I/V curve shows %errors between 50% and 74% from the FRA. Although the irradiance does not affect the open circuit voltage as significantly as temperature, it does cause a slight drop which explain the increase in the error from the I/V curve results. The results show errors that are within acceptable limits from the converter. Figure 9.9 (b) shows similar results except that the series resistance error is larger using the I/V curve, because of the drop in the open circuit voltage with increasing temperature.

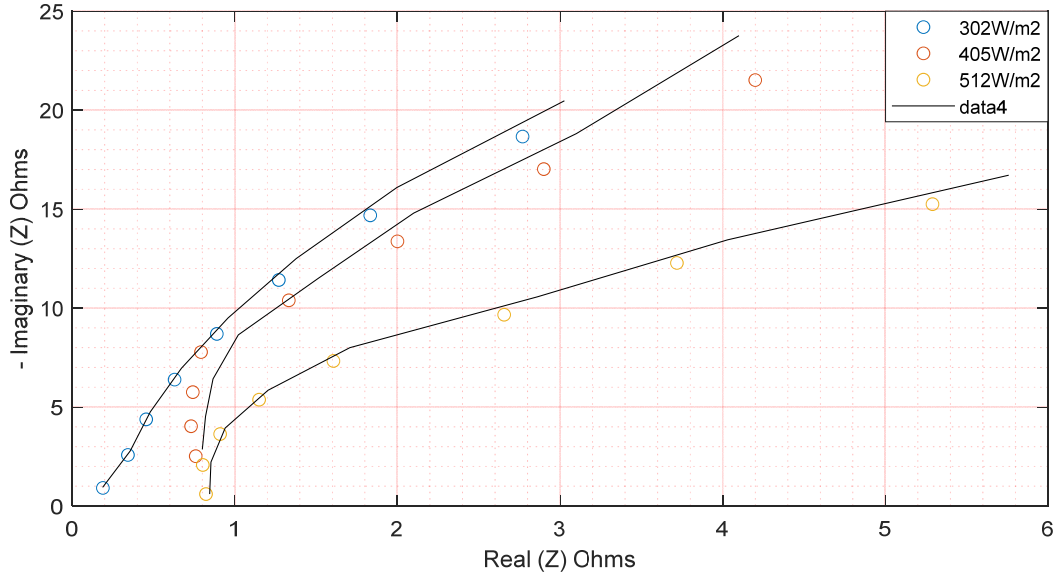


Figure 9.11 Inductor current ripple impedance estimation under low irradiance conditions

Figure 9.10 shows that under high irradiance conditions, more than half of the impedance curve can be characterized using the inductor current ripple. The lowest switching speed used was 20kHz and under these conditions, it exists beyond the maximum imaginary impedance value. This shows that under high irradiance conditions, faster chemical processes within the silicon diode are dominant. Under lower lit conditions, however, the inductor current ripple can only characterize about a quarter of the module's response. Under these conditions the module responds better to lower frequency perturbations as observed from Figure 9.11. Table 9.2 shows that the trends in the parameters are consistent with those observed during Chapter 5. Under good lighting conditions, the series resistance and diffusion information can be fully approximated. The series resistance and diffusion capacitance could be determined from the curve; however, the diffusion resistance was approximated through the method discussed in section 9.7.2; the majority carrier information, however, did not converge.

Table 9.2 Monocrystalline solar module parameter estimation from the inductor current ripple

Parameter	702 W/m^2	616 W/m^2	512 W/m^2	405 W/m^2	302 W/m^2
r_s	$1.51 \pm 0.2\Omega$	$1.1 \pm 0.4\Omega$	$915 \pm 20\text{m}\Omega$	$396 \pm 10\text{m}\Omega$	$786 \pm 320\text{m}\Omega$
r_{np}	$\approx 10 \pm 0.8\Omega$	$\approx 12 \pm 0.7\Omega$	—	—	—
c_{np}	88.42mF	61.22mF	—	—	—
r_{pp+}	—	—	$821 \pm 85\text{m}\Omega$	$1.68 \pm 0.4\Omega$	—
c_{pp+}	—	—	$39.79 \pm 2\text{mF}$	$15.92 \pm 6\text{mF}$	—

Because under lower lit conditions less than half of the curve could be characterized, the diffusion information could be approximated because the peak imaginary response was undefined for the conditions; however, the majority carrier information for both 500 W/m^2 and 400 W/m^2 conditions

converged and could be accurately determined from the curve. The table does show that, with inductor current ripple excitation, the higher the irradiance, the better the characterization; hence under 300 W/m^2 on a single parameter could be approximated. The data acquisition time for the measurements averaged to, 2.30mins for each of the curves; significantly slower than broadband implementation.

9.10 Chapter conclusions

In this chapter, an analysis into the use of a boost converter's inductor current ripple as an EIS excitation signal, for the characterization of a monocrystalline solar module, was undertaken. This approach was in pursuit of a reduced cost of implementing EIS on solar modules through the use of DC-DC power converters, resulting from the solar module's wide impedance bandwidth. The analysis showed that the magnitude of the inductor current's fundamental frequency is independent of the converter's operating point. However, the EIS measurement frequency range is significantly dependent on the operating point of the converter, with higher operating voltages and system inductance offering wider frequency ranges. The inductor's DCR showed significant impact on the expected current ripple magnitude and is desired to be as low as possible to maximize the excitation signal, especially at low load currents. The analysis showed that it is better to design the nominal system for lower operating speeds compared to the maximum required impedance for characterization – this helps the system achieve better performance in its power regulation during the impedance measurements.

The results showed that while there was an error in the parameters extracted, the implementation benchmarks well with the FRA and comparable to DC methods for performance characteristics. There are consistencies between the series resistance measured from the module's I/V curve and the real component of both the FRA and converter; this value was consistently present around the 100kHz frequency component. What is evident is that EIS performs better than the I/V curve at lower resistance values. A range of EIS frequencies within the inductor current ripple's optimal design range, were chosen to ensure that while EIS is implemented, the power is regulated well across the battery load. Because the module response better to higher frequencies under high irradiance conditions, inductor current ripple excitation through the converter is more suited for these conditions; in fact, the series resistance of the module could be characterized under any condition using the inductor current ripple, particularly because it occurs at much higher frequencies, whereas lower light conditions are more suited for EIS measurements taken from an external perturbation signal.

The major drawback to this implementation is that the impedance is measured using a single frequency at a time – contrary to broadband implementation; thus, although the impedance measurements can be measured while the solar is operational, the data acquisition time, for measurements within a single condition, is significantly longer. The method is thus, best suited for single frequency characterization such as, the module's series resistance which consistently appears around the, 100kHz

component; that is, operating the converter at this frequency gives easy access to the module's series resistance across various operating points. This can be of great significance because in Li-ion batteries, the series appears around the, 1kHz component, as such, this frequency is used in industry to fully characterize the Li-ion battery [11]. With continuing research, a similar approach could be adopted for solar modules for optimal cost reduction during frequency characterization.

9.11 References

- [1] R. Choi, J. Shim, and J. Choi, "Audible Noise Reduction in Switched-Mode Power Supplies Operating in Burst Mode," 2021 IEEE Green Technologies Conference (GreenTech), 2021, pp. 528-532, doi: 10.1109/GreenTech48523.2021.00088.
- [2] X. Ke, Z. Hay, and S. Tian, "Kappa Switching DC-DC Converter with Continuous Input and Output Currents Achieving 86.7% Input Ripple Suppression and 16dB Peak EMI Reduction," 2020 IEEE Applied Power Electronics Conference and Exposition (APEC), 2020, pp. 1315-1319, doi: 10.1109/APEC39645.2020.9124150.
- [3] Xue Wenyan, Chen Xiao, and Ma Xikui, "Analysis and reduction of conducted EMI from an AC/DC high power converter," 2005 IEEE International Symposium on Microwave, Antenna, Propagation and EMC Technologies for Wireless Communications, 2005, pp. 503-506 Vol. 1, doi: 10.1109/MAPE.2005.1617959.
- [4] Paulene Govender, Michael J. Brooks, Alan P. Matthews, "Cluster analysis for classification and forecasting of solar irradiance in Durban, South Africa", 2018 Journal of Energy in Southern Africa, vol.29.
- [5] Peña, Rafael & Ospino, Adalberto & Segundo, J. & Rodriguez, Aurelio & Noriega Angarita, Eliana & Muñoz, Yecid. (2020). Economic and Energy Analysis of Small Capacity Grid-connected Hybrid Photovoltaic-wind Systems in Mexico. International Journal of Energy Economics and Policy. 10. 7-14. 10.32479/ijeep.8449.
- [6] M. Shanmugam, B. Bills, M. F. Baroughi and D. Galipeau, "Analysis of carrier transport in dye sensitized solar cells using temperature dependent dark I-V measurements," 2009 34th IEEE Photovoltaic Specialists Conference (PVSC), 2009, pp. 001992-001994, doi: 10.1109/PVSC.2009.5411512.
- [7] A. S. Eeles, R. Gottschalg and T. R. Betts, "Optimising I-V measurements of high capacitance modules using dark impedance measurements," 2016 IEEE 43rd Photovoltaic Specialists Conference (PVSC), 2016, pp. 3693-3697, doi: 10.1109/PVSC.2016.7750366.
- [8] J. Salinger, V. Benda, and Z. Machacek, "Determination of recombination centres in c-Si solar cells from dark I-V characteristics," 2010 27th International Conference on Microelectronics Proceedings, 2010, pp. 149-151, doi: 10.1109/MIEL.2010.5490511.

- [9] Yuancheng Ren, Ming Xu, Jinghai Zhou and F. C. Lee, "Analytical loss model of power MOSFET," in IEEE Transactions on Power Electronics, vol. 21, no. 2, pp. 310-319, March 2006, doi: 10.1109/TPEL.2005.869743.
- [10] M. Marinca, E. Ilies, C. Barbulescu and A. Gontean, "In-circuit ESR Measurement Method for Electrolytic Capacitors," 2021 IEEE 27th International Symposium for Design and Technology in Electronic Packaging (SIITME), 2021, pp. 13-16, doi: 10.1109/SIITME53254.2021.9663652.
- [11] Cuadras, Angel & Kanoun, Olfa. (2009). SOC Li-ion battery monitoring with impedance spectroscopy. 5. 1 - 5. 10.1109/SSD.2009.4956761.

Chapter 10

Conclusions and recommendations

10.1 Concluding remarks

Based on the findings discussed throughout the study, the following concluding remarks were drawn, based on the major themes of the study.

10.1.1 Satisfactory development of a wide bandwidth multiphase converter

The 1st major aim of the study was the development of a wide bandwidth multiphase SMC, as an excitation source, through wide bandgap semiconductor technology. A 2-level multiphase boost configuration was shown to be most suitable to implement impedance spectroscopy, as discussed in Chapter 4. Both simulation and experimental results showed that the maximum achievable bandwidth was only 20% of the switching frequency, to allow for good system stability; this was only 50% of the required excitation bandwidth for the module. However, the frequency spread as discussed in Chapter 5 shows that impedance information between 50kHz and 100kHz has minimal impact on the Nyquist curve – especially under low light conditions.

This is contrary to the hypothesis undertaken that the SiC semiconductor as well as a high frequency driver would improve the system's high frequency performance. Generally wide bandgap semiconductors are designed to achieve better performance at higher system power and voltages; however, the designed system was a low powered low voltage system, which explains why the benefits did not translate to the designed system. However, it was shown through both design and results that the attenuation at excitation frequencies between 50kHz and 100kHz was manageable with good perturbation signal modulation and control; not only that but, excitation frequency spread, as discussed in Chapter 9, showed that this impacts the impedance measurements significantly only in high irradiance conditions; thus, the lower than expected SMC bandwidth was not a terminal factor in the successful implementation of LIS on a solar module.

10.1.2 Satisfactory implementation of perturbation signal amplitude modulation

The 2nd major aim of the study was the adoption of amplitude modulation techniques in both analog and digital perturbation signals; this was in an effort to mitigate frequency induced nonlinearities that result from the induction of an excitation signal through the SMC. These nonlinearities were evident in both an analog chirp and digital PRBS excitation signals, as analysed during both simulation and experimental results; excitation amplitudes at lower frequencies were amplified beyond the point of linear excitation, while higher frequencies were attenuated below the minimum excitation amplitude, thus reducing the signals' effective excitation bandwidths.

Chapters 7 and 8 detailed the amplitude modulation approaches taken both chirp and PRBS signals, and the results showed that, although there was significant complexity in the system design, modulating the amplitude of the perturbation duty cycle signal does improve the amplitude of the induced excitation signals. The modulation does not, however, match the duty cycle signal to the induced excitation signal perfectly in phase as the results showed that impedance estimated from the PRBS duty cycle signal had better phase error at higher frequencies compares to its modulated APRBS signal.

10.1.3 Satisfactory impedance extraction control system

The 3rd major aim of the study was the synthesis of a robust control system for linear excitation signal induction. The study showed that control system required, for the successful implementation of online linear impedance estimation, comprises of 3 architectures that communicate through an embedded control unit; these were, the power regulation, MPPT, and signal injection control architectures. The power regulation was designed such that the current loop can achieve the maximum loop bandwidth required for the induction of the voltage excitation signal, as well as achieving a steady state error that keeps the solar module within its linear range during measurements. The MPPT architecture was designed to keep track of when the module is operating its current or voltage source modes, because as discussed, the impedance measurements were taken while the module operated as a current source. Lastly, through quantitative feedback, the signal injection control was designed to allow for the measurement of impedance, through the SMC, under variable conditions and frequency, to ensure linearity.

The results showed that using QFT control can be used to take impedance measurements accurately under varying operational conditions caused by either irradiance or temperature – as analysed in Chapter 6. the maximum loop bandwidth that could be achieved without affecting system stability was 20% of the converter's switching frequency – this meant that higher frequency components would be attenuated – this was mitigated by the control action adjusting the perturbation amplitudes at these frequencies. Results subsequently showed that the amplitude adjustment improves high frequency power. MPPT control was realised through incremental conductance to ensure fast convergence and less oscillations at steady state under rapidly varying conditions. Quantitative feedback signal injection control

showed how the induced excitation signals could be maintained within the desired range during varying voltage and load conditions, to improve the impedance measurements.

10.1.4 Satisfactory evaluation of crystalline solar modules' impedance as the DUT

In this study, mono and polycrystalline solar modules were analysed as devices under test. The evaluation of the modules was undertaken in real-time. The goal of the evaluation was to assess to what degree can the impedance of the module be characterized using impedance spectroscopy implemented through an SMC. The results showed that the modules respond to frequency perturbations in a manner that depends on its operational, temperature and fault conditions. Under high irradiances the impedance of the module is lower, with however, a higher series resistance – the vice versa was true for lower irradiances. The increasing temperature on the module has the opposite effect, it increased the series resistance and diffusion capacitance, while decreasing the diffusion resistance significantly – this increases the module's charging time constant. The module was shown to respond well to higher frequency perturbations under high irradiance levels – meaning frequency points greater than 10kHz dominated a larger portion of the Nyquist impedance curve – while responding better to lower frequency (frequencies below 10kHz dominating the curve) perturbations at lower irradiance levels. This fact is an important finding because it allows the designer to optimally choose and design excitation signals and excitation frequency ranges.

Under high irradiance, the impedance magnitude of the module decreases, along with the diffusion resistance. The series resistance and diffusion capacitance, however, increasing with increasing irradiance. Temperature had the effect of increasing the series and diffusion resistances, with the diffusion capacitance decreasing, as expected with from the theory of temperature effects of capacitance. These effects were similar to hotspots because of the increasing temperature of the cells. Partial shading had the effect of increasing the impedance magnitude as well and diffusion resistance, with the capacitance decreasing due to the impedance of the bypassing diodes. Thus, detection of partial shading can be characterized by high impedance under high irradiance (which is contrary to these conditions).

With current condition monitoring systems, monitoring this in real-time without taking the system offline, is not possible because I/V curves are used for this; from that, the module would have to be taken offline to trace the curves. However, with the designed system, measuring a high impedance under well-lit conditions could be a good indication of partial shading; it is worth noting that the same logic can be used for detecting hotspots, with the exception that when the module is hot spotted, there would be a significant increase in the module's series resistance as well. Microcracks had the least impact on the module's overall impedance magnitude. What was evident was its increase in the series resistance for reasons discussed during the analysis; with the diffusion parameters not showing significant changes. Hence, the detection of microcracks can be characterized by a significant increase in the series resistance compared to the diffusion parameters. Although the changes in the module's Nyquist

impedance are not accurately definitive for each of the conditions, they do provide a satisfactory indication of occurring issues.

10.1.5 *Satisfactory performance for linear impedance spectroscopy using an SMC on a solar PV module as a characterization technique*

Linear impedance spectroscopy using an SMC on a solar PV module was evaluated as a frequency technique for online fault diagnostics. It was shown through design, analysis, results that the successful induction of an excitation signal containing a broadband spectrum, is dependent upon the DUT's impedance parameters and the SMC's power stage parameters – which are frequency dependent, as well as part of the system's dynamic response i.e., the relationship between the injected perturbation duty cycle signal and the induced excitation signal. Results showed that the low pass filtering characteristics of both the DUT's and SMC's responses, led to high signal gains at low frequencies and high attenuation at higher frequencies.

Impedance measurements showed that partial shading and hot spots were the most detectable solar module faults using impedance spectroscopy – showing wider changes in the impedance, while micro cracks did not show significant changes in the Nyquist impedance. However, the compared to conventional PV monitoring systems such as I/V curves, the major limitation of converter-based BIS is the cost associated with building such a system. This is largely due to the processing power required to process the impedance information. Conventional systems monitor DC signals which do not require as much processing capabilities.

10.1.6 *Relevance of impedance extraction system to industry applications*

Based on the analyses and results, the impedance of the module could be related to some of its operational conditions – namely, irradiance and temperature conditions, partial shading, hotspots, microcracks; thus, if integrated properly, the system's impedance could be used for some online diagnostics of a solar module under such conditions. However, the application still requires further research in key areas for solar system integration.

The first area would be cost; as mentioned during the theory development, residential solar systems are monitored on a string level, because this provides good cost-vs-benefit margins. The proposed system on the other hand provides module level diagnostics. Although this would provide better diagnostics, the cost of integrating such a system for each module in a system would be large; thus, because this type of system design is still in its infancy, the cost-vs-benefit it possesses still has a significant gap.

The second area of consideration would be whether the technique could be adopted to measure string level impedance and have capacity to decouple this impedance into individual modules within the string. Based on current research, Huawei is working a device to measure the I/V curve of a string and then decouple that curve into individual module I/V curves; this type of technique can be adopted for the frequency-based technique for have a better cost-vs-benefit margin.

10.2 Study recommendations

Based on the conclusions drawn, the following recommendations were made.

10.2.1 *Assess other module types.*

The cost limitation and the Covid-19 pandemic limited the assessment of other module types for characterization; however, to fully critique the implementation of LIS on solar modules, a larger variation in the module types is recommended. The variation in the impedance frequency distribution on the monocrystalline module showed interesting results – results which could be useful in the characterization of other module types.

10.2.2 *Assess other conditions and faults.*

The conditions of assessment for this study were limited due to the Covid-19 pandemic, as mentioned in the objectives. However, to create a more robust system, other solar module conditions such as, PID, ageing/degradation with time, accelerated stress tests: thermal cycling, humidity freeze, and damp heat, should be assessed. These conditions are important because they are common solar module testing; more specifically, the assessment of ageing on a module. The assessment of PID should be of great interest because, based on available research on the subject, this type of fault takes months (years in some cases) to detect with current condition monitoring systems; thus, research into whether there are visible impedance changes to the module as a result of PID should be conducted, especially because this frequency-based technique can detect impedance changes on a much smaller scale compared to DC techniques.

10.2.3 *Investigate alternate processing units.*

At the inception of this study, several lightweight and cheaper processing units such as, the Arduino family and C2000 microcontroller family, were investigated for high frequency signal processing; however, there were significant issues discovered relating the processing of information at high frequencies, issues with parallel sampling of the ADC, and issues with simultaneous acquisition and generation of control signals, particularly through a suitable software interface. The NI DAQ was subsequently chosen as the unit of choice because it did not suffer the same drawbacks as the aforementioned units. Thus, a processing-unit-specific study must be performed, focusing specifically on high frequency data acquisition of microprocessors and microcontrollers, to source a cheaper and light-weight processing unit.

10.3 Future work recommendations

Based on the study undertaken, there were several works (outside the scope of this study) recommended for future work, to further advance the field of electrochemical impedance spectroscopy, especially on solar modules. These are detailed below.

10.3.1 Simultaneous Impedance estimation of solar module and battery storage

This study showed that the injection of a single duty cycle perturbation signal induces an input current excitation signal, as well as an output voltage excitation signal. This study, among others have used the induced input current excitation signal; while several research has implemented impedance estimation using the output voltage excitation signal, while charging a Li-ion battery. From this, it is evident that there exists a possibility to perform impedance measurements on both solar module and load battery using a single perturbation signal. In solar and storage hybrid systems, solar modules are configured to store excess energy in the batteries, through a power optimizer of sorts. With this setup, an impedance measurement device, integrated into the power optimizer, can be designed to implement condition monitoring of both solar and battery. The successful execution of the technique implemented in this manner would rest on a solid understanding of both solar module and battery impedance behaviour, and most importantly, a perturbation duty cycle signal designed to accommodate the impedance design requirements of both devices – particularly, the excitation frequency and excitation amplitude. The major challenge to this type of implementation is that, when using an SMC as the excitation source of choice, the designed must ensure that the SMC has continuous input and output currents to observe linearity in the impedance measurements of both devices; the continuity in SMCs' input and output currents was discussed in detail in Chapter 4. There are SMCs that designed to achieve continuity in both input and output currents, and the evaluation of these devices for excitation signals injection would need to be critiqued in relation to their power regulation capabilities. The other challenge to this implementation would be that there are 2 kinds of power optimizers: MPPT and PWM; the characteristics of each would need to be fully evaluated, to choose the most optimal type of charging algorithm. The last factor keep in mind is the effect of the RHP zero on the voltage loop of a boost type SMC, whose impact on the system design was discussed thoroughly in Chapter 4; this will significantly impact measurements when implementing LIS on a battery while it charges from the module. This is because the location of the RHP zero is likely to occur within the battery's bandwidth.

10.3.2 Partial shading assessment through the impedance characterization of the bypass diodes

As discussed in Chapter 5, the assessment of the module's partial shading condition, relies on the understanding of the module's bypass diode – more specifically, its frequency response. When using the FRA to perform impedance spectroscopy on the body diode of the module, it was observed that potentiostatic characterization is preferred over galvanostatic. Galvanostatic characterization failed in the excitation of the diode's equivalent $P-N$ junction capacitance – it only provided characterization of the junction's resistive impedance. This is largely because of the diode's activation voltage; current excitation does not provide the required voltage bias for the diode to conduct. Potentiostatic characterization, however, provides the required voltage bias, as well the characterisation of the diode's full junction characteristics. This is thus, another major reasons why potentiostatic excitation is preferred for solar

modules, that has not been fully identified in literature. Another aspect to consider, is the failure of the bypass diode – more specifically, its response during faulty and normal working conditions. All the above shows that there needs to a fully dynamic assessment of a solar module under partial shading considerations. If a bypass diode fails, the technique has the potential to fully detect the failure, if investigated accordingly.

10.3.3 Investigate the use of a square wave frequency sweep signal for reduced signal processing power requirements

It was recommended that when making use of the chirp excitation signal for wide bandwidth devices, such as the silicon-based monocrystalline solar module, an FPGA should be considered to improve resolution at higher excitation frequencies. However, because of the cost issue discussed, a square wave frequency sweep signal should be considered. The number of samples required for digital signals is smaller than those required for analog signals, as discussed during experimental verification. The lower number of samples will ensure a higher sampling rate when using devices such as the NI DAQ; the higher sampling rate will, in turn, improve the resolution at higher excitation frequencies. There is, however, not much work on the square wave sweep signal as a device excitation signal; one of the reasons for this could be that the chirp sweep signal is a well-researched signal, with a large volume of existing code in multiple software for its generation; from this, it makes it easier to implement. The other reason is that parametrising a sinusoidal signal with time i.e., the analog chirp signal, is easier than formulating a parametric digital sweep signal. Perhaps the most crucial issue would be that, as a square wave sweep signal, there may be a large potential unwanted harmonic within the signal and hence, lower SNR. However, these aspects would be investigated as part of the study into the use of the square wave sweep as an excitation signal.

10.3.4 Investigate the use of a multiphase converter to control inductor current ripple magnitude during device excitation

During the assessment of the converter's inductor current ripple for device excitation, a single-phase boost converter was used for the excitation to maximise the inductor current ripple. However, as established, when designing the converter there are certain trade-offs between maximizing the inductor current ripple magnitude and ensuring optimal power regulation. Because of this, introducing another dynamic layer to control the current ripple magnitude without impacting the power regulation, is desired; this layer would be a multiphase SMC. The multiphase converter can regulate the current ripple magnitude with minimal impact on the power regulation. The more phase there are in the converter, the higher the degree of current ripple regulation with little trajectory change in the duty ratio. This is important because a larger change in the duty ratio will cause nonlinear operation. Another issue to consider is the dynamic variability in a system with an increased number of storage elements, which could

cause stability issues in the control system. Lastly, there is, the issue of losses that increase with number of phases.

10.3.5 High frequency impedance prediction algorithm

As discussed throughout the study, accurate high frequency impedance characterization of the solar module requires expensive processing hardware. In Chapter 9, an alternate approach was undertaken in a cost-reduction effort; however, it was limited to high-frequency measurements only. The question thus becomes, is it possible to design an online impedance extraction system by implementing impedance spectroscopy such that high-frequency information can be modelled or predicted by analysing the measured low-frequency information or vice-versa? An externally generated duty cycle perturbation signal would be used in the case of high-frequency information modelling from low-frequency information. In Chapter 5, the frequency distribution of the impedance information was analysed; it was shown that the definition of “high-frequency” information is condition dependent; therefore, a more detailed chemical-level analysis would have to be undertaken to fully characterize the nature of the module’s response to frequency perturbation in varying irradiance and temperature; the understanding of this nature, will help design a prediction system that translates into usable results that can be interpreted for a better monitoring system.

Appendices

Appendix A Experimental setup

This appendix includes photos of the experimental setup and its components.

A 1. Frequency Response Analyser setup

The FRA was setup as shown in the laboratory.



Figure 10.1 AUTOLABPGSTAT302N frequency response analyser

A 2. Power optimizer prototype

This is the lab setup for the multiphase boost converter.

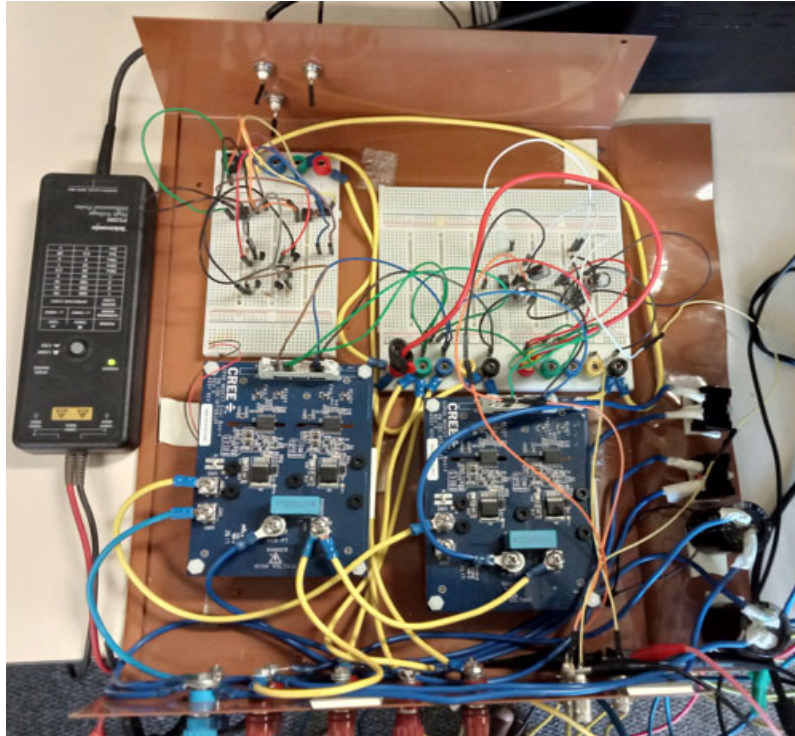


Figure 10.2 Multiphase boost power optimizer prototype top view

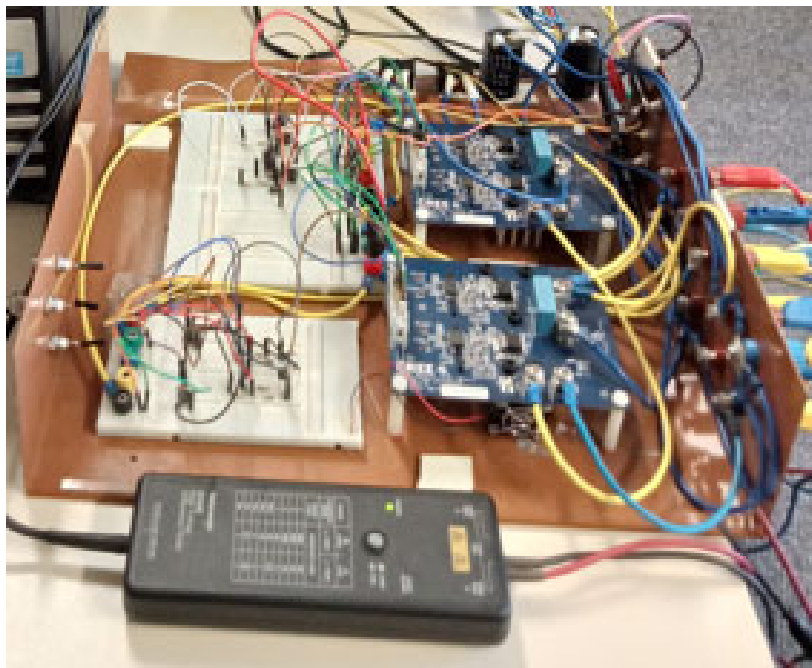


Figure 10.3 Multiphase boost power optimizer prototype side view

A 3. MOSFET power driver

The following set of pictures give a detailed view of the CRD-5FF0912P, SiC MOSFET High-Frequency Evaluation Board for 7L D2PAK.

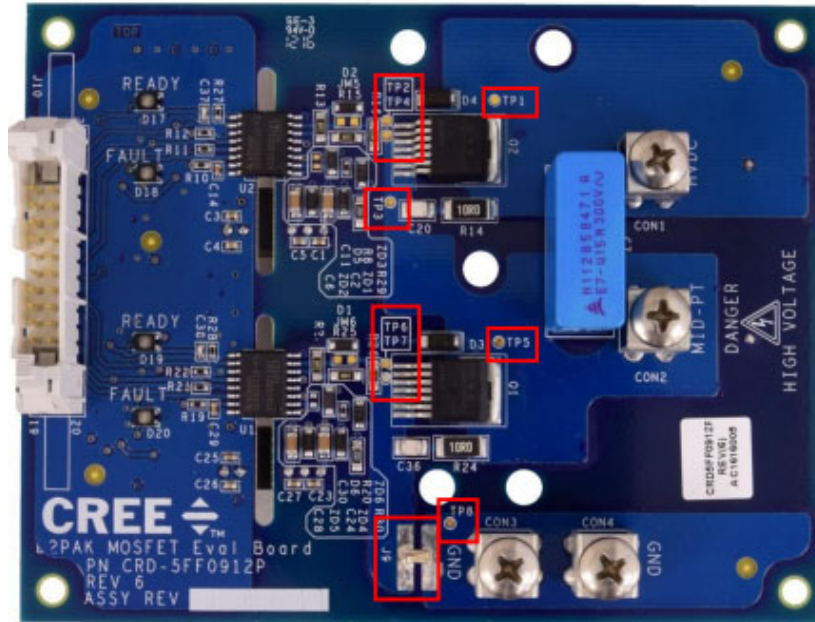


Figure 10.4 CRD-5FF0912P, SiC MOSFET High-Frequency Evaluation Board for 7L D2PAK top view.



Figure 10.5 CRD-5FF0912P, SiC MOSFET High-Frequency Evaluation Board for 7L D2PAK side view.

Appendix B MATLAB Simulation Models

B 1. Boost SMC models

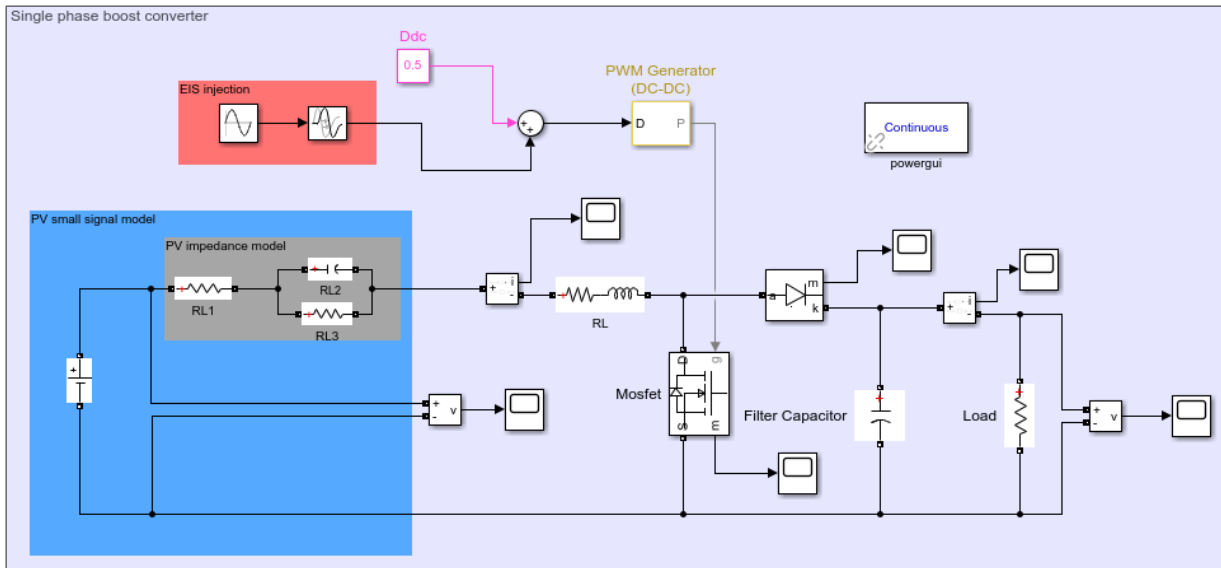


Figure 10.6 Single phase boost converter model showing impedance estimation

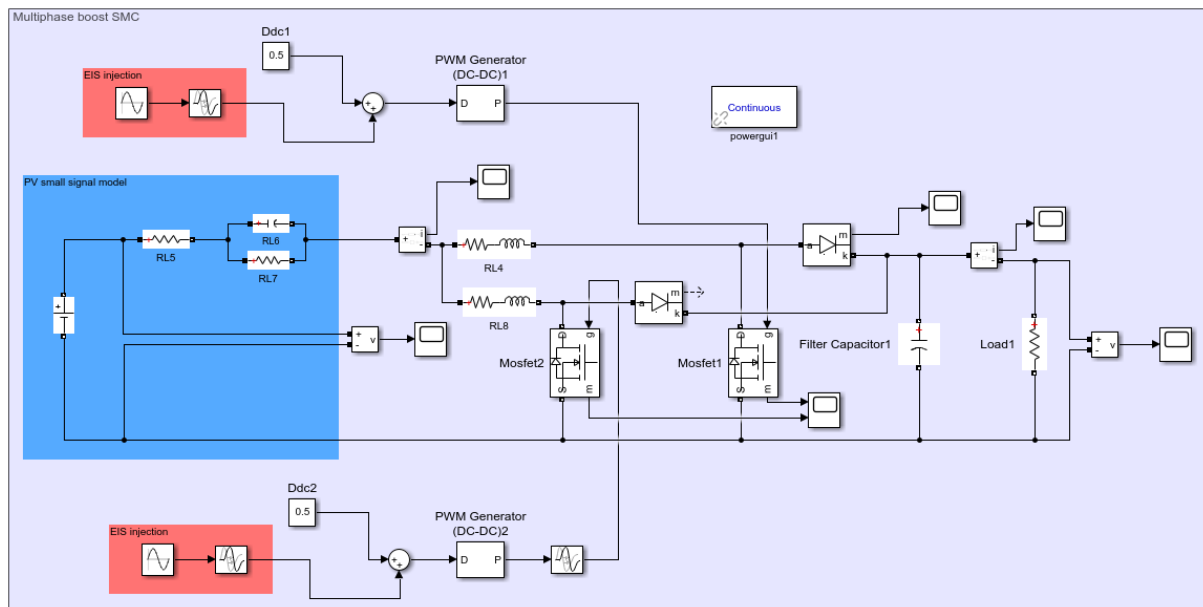


Figure 10.7 Multiphase phase boost converter model showing impedance estimation

B 2. Load control models

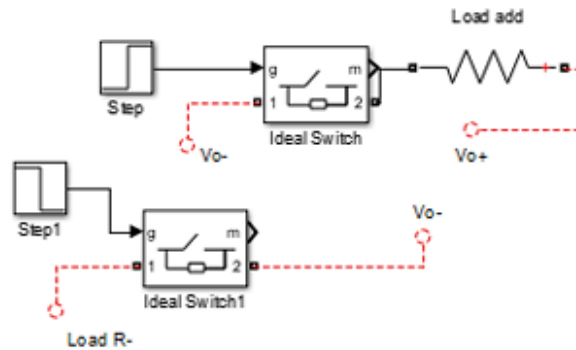


Figure 10.8 Load addition and removal model

B 3. Power regulation control model

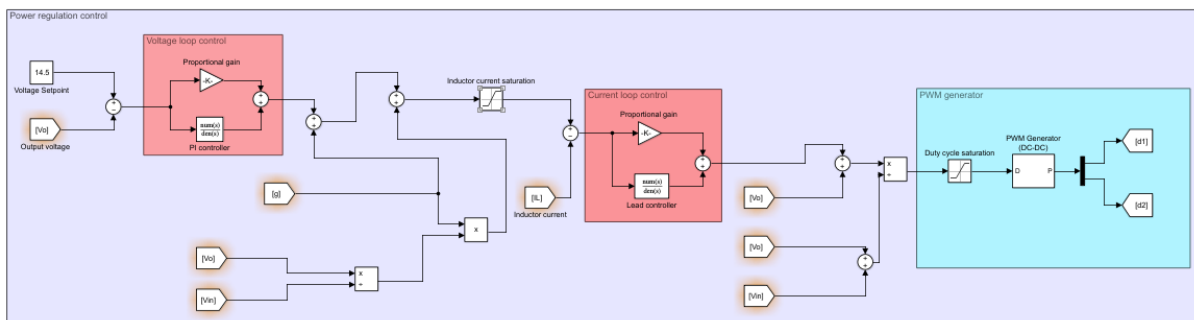


Figure 10.9 Power regulation control MATLAB graphical model.

B 4. MPPT control model

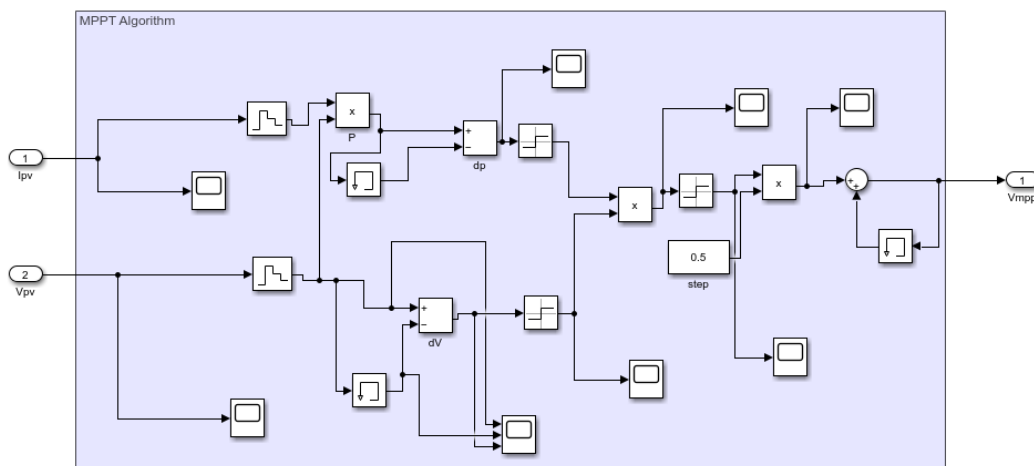


Figure 10.10 MPPT control MATLAB graphical model.

B 5. Full 3-tier control model

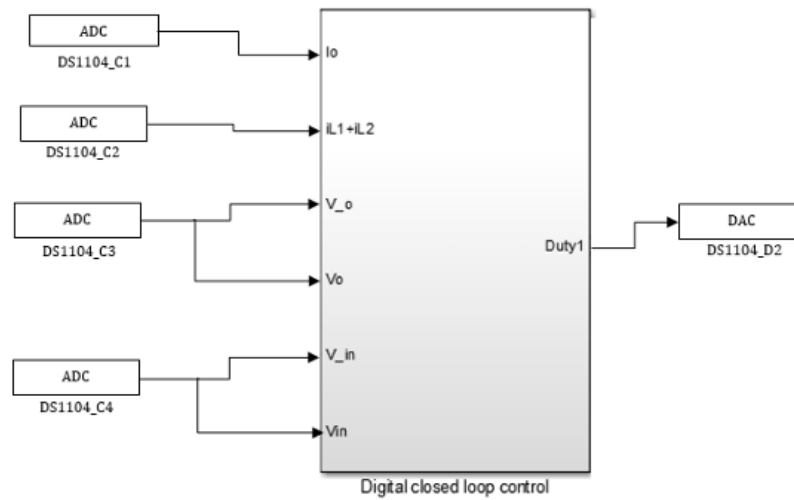


Figure 10.11 Closed loop MATLAB Simulink digital control for the practical system

Appendix C C-Code for QFT control, signal processing, and excitation signals

C 1. Chirp function

```
function [ z_re, z_im ] = chirp_imp( x, y)
%UNTITLED2 Summary of this function goes here
% Detailed explanation goes here

%% simulation params
fs = 2000000; % sampling freq
t = 0:1/fs:1; % simulation time

f_0 = 10; % initial freq
f_t = 100000; % target freq
t_t = 1; % target time from block

%% params
tbw_start = 20; tbw_end = 20;
gamma = 6;

freqs = [4 5 6 8 10 12 14 16 20 26 32 40 50 64 72 80 90 100 128 150 175 200 230
256 300,...
```

```

    350 400 450 512 600 700 800 900 950 990, 1000, 2000, 5000]; % maybe make an
argument
%maxmin = [min(freqs), max(freqs)];
%% the chirp nuances
%beta = (f_t - f_0)/t_t^2;    % beta value

beta_log = (f_t/f_0)^(1/t_t);

freq_times = zeros(length(freqs), 1);
for i = 1:length(freqs)
    %freq_time = sqrt((freqs(i) - f_0)/beta); % consider vectorizing
    freq_time = log(freqs(i) / f_0)/(log(beta_log));
    freq_times(i) = freq_time;
end

%% get the coefs with cwt
[wt_x,f_x, wt_y, ~] = wavlt(x, y,t, gamma, tbw_start, tbw_end, fs);

%% locating coeff
x_coeffs = zeros(length(freqs), 1); %use zeros
y_coeffs = zeros(length(freqs), 1);

for j = 1:length(freqs)
    freq = freqs(i);
    freq_time = freq_times(j); % time when the freq occurs
    [~, f_idx] = min(abs(freq - f_x)); % get freq index
    [~, t_idx] = min(abs(freq_time - t)); %

    % remember x is voltage so the current y is sort of the reference
    freq_loc_x = wt_x(f_idx,:); % localize in freq (for all time)
    freq_loc_y = wt_y(f_idx,:);

    x_coeff = freq_loc_x(t_idx); % hmm how do I localize this?
    %x_coeff = max(freq_loc_x);
    y_coeff = freq_loc_y(t_idx); % localize in time

    x_coeffs(j) = x_coeff;
    y_coeffs(j) = y_coeff;

```

```

end

imps = x_coeffs./y_coeffs;
z_re = real(imps);
z_im = imag(imps);
%imp_data = [z_re, -z_im];
figure(3)
plot(z_re, -z_im, '-*');
semilogx(freqs, abs(imps) )
grid('on');

% time = datestr(datetime);
% time = strrep(time, ':', '-');
% filename = strcat(time, string(-fs), string(-maxmin(1)), string(-maxmin(2)),...
%     string(-gamma), string(-tbw_start), '.csv');
% csvwrite(filename, imp_data);
end

```

C 2. Wavelet transform

```

function [wt_x,f_x, wt_y, f_y] = wavlt(x, y, t, gamma, tbw_start,...
    tbw_end, fs)

nv = 48;    % numberof Voices
tbws = tbw_start:2:tbw_end;
gammas = repmat(gamma, [length(tbws), 1]);
wvparams = [gammas tbws'];

%% absolutely useless stuff I've done here loop should be in main
for i = 1:size(wvparams,1)

    wvparam = wvparams(i,:);
    %% for x
    [wt_x, f_x] = cwt(x,fs, 'VoicesPerOctave', nv,...
        'WaveletParameters',wvparam);
    xrec = icwt(wt_x, 'VoicesPerOctave', nv,...
        'WaveletParameters',wvparam);

    figure(i)

```

```

        plot(t, xrec);

        %% for y
        [wt_y, f_y] = cwt(y,10000, 'VoicesPerOctave', nv,...
            'WaveletParameters',wvparam);
        yrec = icwt(wt_y, 'VoicesPerOctave', nv,...
            'WaveletParameters',wvparam);
        figure(i+1)
        plot(t, yrec)
    end

end

```

C 3. Convolution

```

n = -20:20;
Ppv(); %Converter dynamic system
dc (); %Chirp wavelet transform
N = length(Ppv);
M = length(dc);
Ny = N + M -1;
y = zeros(1,Ny);
for i = 1:N
    for k = 1:M
        y(i+k-1) = y(i+k-1) + dc(k)*Ppv(i);
    end
end
m = 0: Ny-1;
% user defined
figure (1)
stem(m,y,'linewidth',3,'color','m')
grid;
a = title('Output of an LTI System y(n)');
set(a,'fontsize',14);
a = ylabel('y(n)');
set(a,'FontSize',14);
a = xlabel('n');
set(a,'FontSize',14);
% Using matlab built in function
figure (2)
y2 = conv(Ppv,dc);
stem(m,y2,'linewidth',3,'color','r')
grid;

```

```

a = title('Output y(n) using conv(x,h)');
set(a,'fontsize',14);
a = ylabel('y(n)');
set(a,'FontSize',14);
a = xlabel('n ');
set(a,'FontSize',14);

```

C 4. Single frequency impedance extraction

```

%%Update Fs,N_priod and freq
Fs=200000;
freq=10;
t_period = 1/freq; % signal period
n_samp_period = t_period*Fs; % sample period of signal (samples in one period)
n_period=3;
N =int16(n_samp_period*n_period);

%%Update Fs,N_priod and freq
Ts=1/Fs;
start_eis=0.2/Ts;
n_period=3;
%%Slicing the signal,to remove the transient state
volt= voltage(start_eis+1:end) - mean(voltage(start_eis+1:end));
curr = current(start_eis+1:end) - mean(current(start_eis+1:end));
% volt=filter(lpFilt,volt);
% curr=filter(lpFilt,curr);

%%Perform the Fast Fourier Transform
v_fft = fft(volt);
i_fft = fft(curr);

%%Index my fast fourier arrays, to obtain the EIS impedance
imp_cmplx = v_fft(n_period+1)/(i_fft(n_period+1));
z_re = real(imp_cmplx); % the impedance gotten is negative (already)
z_im = -imag(imp_cmplx);

```

C 5. QFT control

C5 i. Plant Definition

```

k=qpar('k',,,,);
a=qpar('a',,,,);
z=qpar('z',,,,);
wn=qpar('wn',,,,);

num = [k*wn*wn k*wn*wn*a];
den = [1 2*z*wn wn*wn];
P = qplant(num,den);
k = qpar('k',2,2,5,8)
exp = k*wn*wn
den = [1 2*z*wn wn*wn]
P = qplant(num,den)
w_nom = logspace(-2,2,200);
P.cnom(w_nom)
figure, P.nominal.bode()
figure, P.bodcases()
pars = [1 3; 2 5; 8 4; 0.3 0.3];
figure, P.bodcases(pars)

```

C5 ii. Templates

```

w = [0.2 0.5 1 3 5 7 10 20 50];
P.ctpl('recedge',w)
P.showtpl()

```

C5 iii. Specifications

```

spec1 = qspc('odsrs',w,6)
spec2 = qspc.rsrs([1.2 0.2],10,1.5,[],logspace(-1,2),2.85,3.1)
spec1.show();
des = qdesign(P,[spec1 spec2])
des.cbnd('odsrs')
des.showbnd('odsrs')
des.cbnd('rsrs')
des.showbnd('rsrs')

```

C5 iv. Loop Shaping

```

h = des.showbnd('odsrs',[],[2 10 1e2 5e2 7e3 10 50e3 100e3]);
des.showbnd('rsrs',h,[0.2 0.5 1]);

% define the compensator G(s):
s = zpk(0,[],1);
set(s,'DisplayFormat','Frequency')
G = ();

des.loopnic(G) % plot G*Pnom on NC
ngrid

```

C5 v. Prefilter

```

spec2.show('freq');
des.clmag(G,1)
ylim([-55 10])
F = 1/(...*s/..s^2/..)
spec2.show('freq');
des.clmag(G,F)
ylim([-55 10])

```

C5 vi. Design Validation

```

L = series(P,G);      % open loop
S = feedback(L,1);    % closed loop from d to y (sensitivity)
T = series(S,series(L,F)); % closed loop from r to y
spec1.show; hold on      % show the specs.
pgrid = P.templates(6).parameters; % take parameters used for biggest template
S.bodcases(pgrid,w_nom,'showphase',0) % plot magnitude response
axis([0.2 50 -20 10])
spec2.show('freq'); hold on % show the specs.
T.bodcases(pgrid,w_nom,'showphase',0) % plot magnitude response
axis([0.01 100 -55 10])

```