

Modelling and Analysis of a Two-Input Cuk-Type DC-DC Buck-Boost Converter for Space Applications

Vankara Mounika¹, CH V V S Bhaskara Reddy², B Amarendra Reddy³, Ch. Sruthi Varma⁴

¹ MTech Student, Department of Electrical and Electronics Engineering, Andhra University, Andhra Pradesh, India

² Professor, Department of Electrical and Electronics Engineering, Andhra University, Andhra Pradesh, India.

³ Associate Professor, Department of Electrical and Electronics Engineering, Andhra University, Andhra Pradesh, India.

⁴ Research Scholar, Department of Electrical and Electronics Engineering, Andhra University, Andhra Pradesh, India.

Abstract - This paper reports the modelling, design and control-interaction study of a Cuk-derived two-input buck-boost converter that merges the power of two independent DC sources in a single conversion stage. The converter keeps the main strengths of the classical Cuk topology, namely a non-pulsating input current and a small output-current ripple, and additionally lets both sources supply the load at the same time. Input sources of 24 V and 12 V are used here to produce a regulated output of -52 V. A state-space averaged model is built from the three switching intervals, from which an expression for the voltage gain as a function of the two duty ratios is obtained; this expression is cross-checked with the volt-second balance principle. Using the resulting gain relationship, the passive elements are sized for duty ratios of 0.5 and 0.7 at a switching frequency of 50 kHz. MATLAB/Simulink simulations then confirm that the output voltage agrees well with the predicted value. The averaged model is next linearized about the chosen operating point, which yields small-signal transfer functions from the duty-ratio inputs to the converter outputs. The interaction between the two control loops is quantified by applying the Relative Gain Array (RGA) to the resulting transfer-function matrix. The diagonal RGA entries are about 0.997, which points to weak coupling and almost decoupled control channels. Consequently, each duty ratio can be regulated by its own single-loop controller, and no more elaborate multivariable control scheme is needed.

Symbol	Explanation
C_1	Coupling capacitor (Cuk energy-transfer capacitor)
C_2	Output capacitor
R	Load resistance
r_{L1}, r_{L2}	Parasitic resistances of the inductors
r_{C1}, r_{C2}	Equivalent series resistances of the capacitors
G_v	Voltage gain of the converter
SSA	State-space averaging
V_{g1}	Supply voltage of source 1 (12 V)
V_{g2}	Supply voltage of source 2 (24 V)
V_0	Regulated output voltage (-52 V)
TF	Transfer function

Key Words: Dual-input DC-DC converter, Cuk converter, Buck-boost converter, State-space averaging, Small-signal modelling, Voltage gain, Relative Gain Array, Control interaction

NOMENCLATURE

Symbol	Explanation
d_1 and d_2	Duty ratio of switches S_1 and S_2
f_s	Switching frequency (50 kHz)
L_1 and L_2	Input inductors for sources 1 and 2

1. INTRODUCTION

The growing share of renewable and distributed generation has raised the need for power-conversion systems that can pool several independent energy sources into one regulated output. Photovoltaic arrays, fuel cells, batteries and supercapacitors differ in voltage level, current behaviour and dynamic response. Traditionally, each source is served by its own converter stage, which raises the count of semiconductor switches and passive elements and also adds to the size, cost and control burden of the whole system. Multi-input converters (MICs) address this by letting several sources feed one shared

conversion stage, thereby lowering the number of components and making better use of the system [1].

Within the family of MICs, Cuk-derived converters stand out for their non-pulsating input current, small input-current ripple and built-in ability to step the voltage both up and down. These features suit applications in which the quality of the source current and a wide conversion range both matter. A new dual-input Cuk-based converter was studied in terms of its modelling, operating behaviour and design, showing that Cuk-derived structures can combine two independent sources in one conversion stage [2]. The power-flow coupling in double-input Cuk-buck DC-DC converters has likewise been examined, and it was shown that the two sources affect each other through the shared energy-transfer path [3]. Because of this coupling, accurate mathematical models and carefully chosen control schemes are especially important for dual-input Cuk-type converters.

Research on multi-input converters rests on a large body of work covering topology generation, synthesis and operating characteristics. One proposed multiple-input DC-DC topology merges different energy sources and removes the need for separate conversion stages [4]. A systematic method for deriving multi-input DC-DC converters from ordinary converter structures was developed afterwards [5]. Feasible multiple-input topologies have also been identified in a structured way from practical switch-diode arrangements [6]. Other synthesis approaches have been reported that generate multiple-input DC-DC structures from different combinations of conventional converter cells [7]. The distribution of power among the individual sources in multiple-input buck-boost converters has been treated as a power-budgeting problem [8]. Double-input PWM DC-DC converters were built for systems with sources at different voltage levels, laying the groundwork for coordinated use of dissimilar energy sources [9]. How multiple-input DC-DC converters behave under varying source and load conditions has also been examined [10].

Multi-input converters have since found use in renewable-energy and transportation systems. One multi-input DC-DC converter built around a multiwinding transformer let several sources share a common conversion structure in renewable-energy applications [11]. Multiple-input power converters were also studied for hybrid vehicles, where different energy sources have to be coordinated to meet the load demand [12]. A zero-voltage-switching multi-input converter was introduced to raise efficiency, showing that switching losses in multi-source systems can be cut [13]. A further high-efficiency DC-DC converter with two input power sources was designed to improve source utilization and power transfer [14].

Multiport converter architectures have also drawn strong interest for spacecraft and energy-storage uses. A three-port DC-DC converter was modelled and controlled for satellite applications, giving one integrated interface among sources, storage elements and loads [15]. Its dynamic response under different operating conditions was examined in a later study of a three-port DC-DC converter for space use [16]. A family of three-port half-bridge converters was introduced for stand-alone renewable power systems, which shows that multiport structures suit systems that must manage sources and storage together [17].

With more input sources and ports, the interaction among the individual power-processing paths becomes a significant factor in control design. Unlike a single-input converter, a multi-input converter has inherently coupled multivariable dynamics, so a change in one control variable can influence several outputs or power paths. For this reason, decoupling control has been studied for triport DC-DC converters in order to weaken the interaction among the energy interfaces and improve independent control [18]. Transformer-coupled multiport zero-voltage-switching bidirectional converters were proposed to deliver bidirectional power over a wide input-voltage range [19]. Multiport structures have also been developed to integrate distributed generation and storage inside solid-state transformer architectures [20]. Taken together, these works show that dynamic coupling and control interaction must be considered when designing multi-input and multiport power-conversion systems.

In spite of this progress, little attention has been paid to a systematic assessment of control interaction in a two-input Cuk-type buck-boost converter in particular. Earlier work has concentrated mainly on topology, steady-state behaviour, power-flow characteristics, synthesis and dynamic modelling. Since both duty ratios act on the output through one shared energy-transfer network, designing separate controllers from the individual control-to-output transfer functions alone may fail to capture the true multivariable behaviour. A quantitative measure of the interaction between the two control channels is therefore needed to establish how strongly they are coupled and whether independent single-loop control is adequate.

To this end, the present paper carries out a systematic modelling, design and control-interaction study of a two-input Cuk-type buck-boost converter for multisource power applications. The converter is first described by state-space equations for each of its three switching modes, and an averaged state-space model is then constructed. The steady-state voltage gain is derived and cross-checked with the volt-second balance principle. For the chosen operating conditions, the passive components are sized and the converter is validated in

MATLAB/Simulink. Next, the averaged model is linearized about the operating point, giving the small-signal control-to-output transfer-function matrix for the two duty-ratio inputs. Finally, Relative Gain Array (RGA) analysis measures the interaction between the two control channels and indicates whether independent single-loop control is suitable.

The main contribution is an integrated analytical framework linking switching-state modelling, averaged modelling, steady-state gain verification, small-signal analysis and multivariable interaction assessment for a two-input Cuk-type buck-boost converter. It gives a clear picture of how the two input channels are dynamically coupled and lays the basis for later controller design and decoupling strategies in multi-source systems, especially where compact, efficient and dependable power processing is required.

2. MULTI-INPUT CUK-TYPE DC-DC BUCK-BOOST CONVERTER

This converter merges several input sources and delivers a single regulated output. Its operating principle comes from the well-known Cuk converter, which has a continuous input current and low output ripple. The proposed circuit extends the Cuk principle by using two

input sources in place of one. Each source has its own inductor, so both input currents flow continuously and the stress on the sources is reduced [2]. The coupling capacitor separates the input and output sides and serves as the intermediate energy-storage element. The switches that regulate the sources operate with duty cycles of d_1 and d_2 , while the diodes provide freewheeling paths during switching. The output side consists of a capacitor and a load resistor, and the output voltage is inverted.

To characterize the design and control behaviour of the circuit, state-space averaging is used to obtain the voltage gain and the transfer functions. The RGA analysis shows that the interaction between the control loops is small, so each input can be controlled independently [6],[7]. Considering these properties together with two-input operation, efficient energy transfer and good controllability, the converter is well suited to demanding conditions such as aerospace and space power generation [8], [1].

2.1 Modes of Operation

In each switching period, the converter goes through three operating modes, for the operating point $d_1 > d_2$.

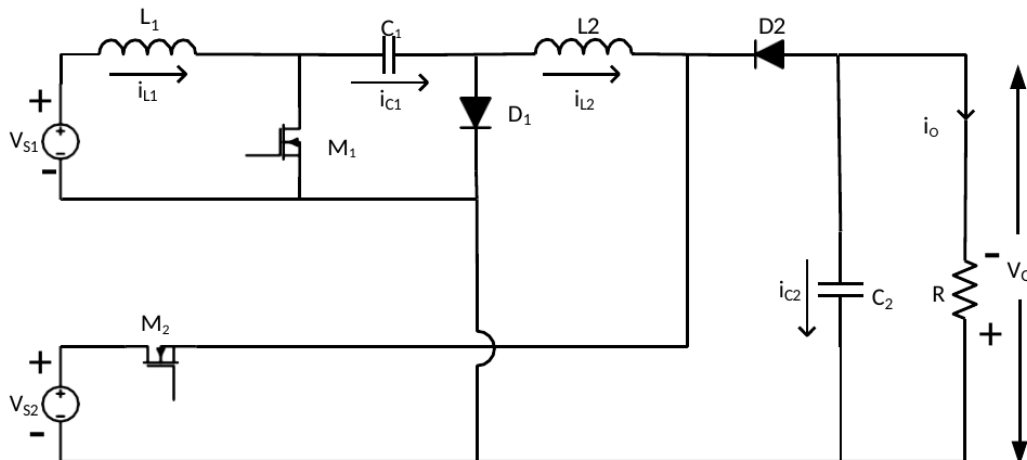


Fig -1: Circuit diagram of the multi-input Cuk-type converter

Table -1: Modes of operation

Operating mode	Time interval	Sources feeding load R	Conducting switches
Mode 1	$0 < t < d_2 T_s$	$V_{g1} + V_{g2}$	S_1, S_2
Mode 2	$d_2 T_s < t < d_1 T_s$	$V_{g1} + V_{g2}$	S_1
Mode 3	$d_1 T_s < t < T_s$	V_{g2}	None

2.1.1 Mode 1 ($0 < t < d_2 T_s$)

Fig -2 shows the circuit topology during Mode 1 operation. Here the diodes D_1 and D_2 are off, while the switches S_1 and S_2 are on. The input voltages V_{g1} and V_{g2} ,

respectively, charge the inductors L_1 and L_2 . Kirchhoff's laws are used to write the circuit equations, with the inductor currents and capacitor voltages taken as the state variables (the current paths are shown in Fig -2). The dynamics of this mode then follow as:

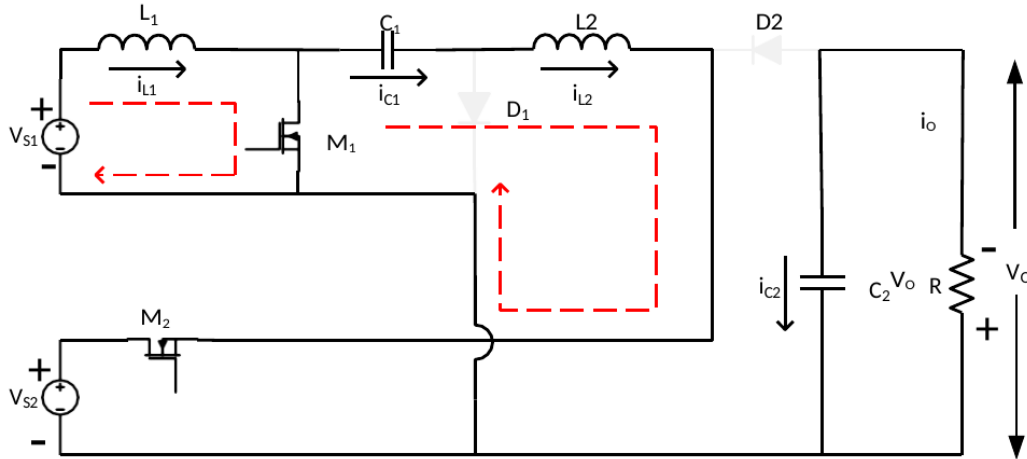


Fig -2: Current-flow paths during Mode 1 operation ($0 < t < d_2 T_s$)

For L_1 :

$$\frac{di_{L1}}{dt} = -\frac{r_{L1}}{L_1} i_{L1} + \frac{1}{L_1} V_{S1} \quad (1)$$

For C_1 :

$$\frac{dV_{C1}}{dt} = \frac{1}{C_1} i_{L2} \quad (3)$$

For L_2 :

$$\frac{di_{L2}}{dt} = -\frac{r_{L2} + r_{C1}}{L_2} i_{L2} - \frac{1}{L_2} V_{C1} + \frac{1}{L_2} V_{S2} \quad (2)$$

For C_2 :

$$\frac{dV_{C2}}{dt} = -\frac{V_{C2}}{(R + r_{C2})C_2} \quad (4)$$

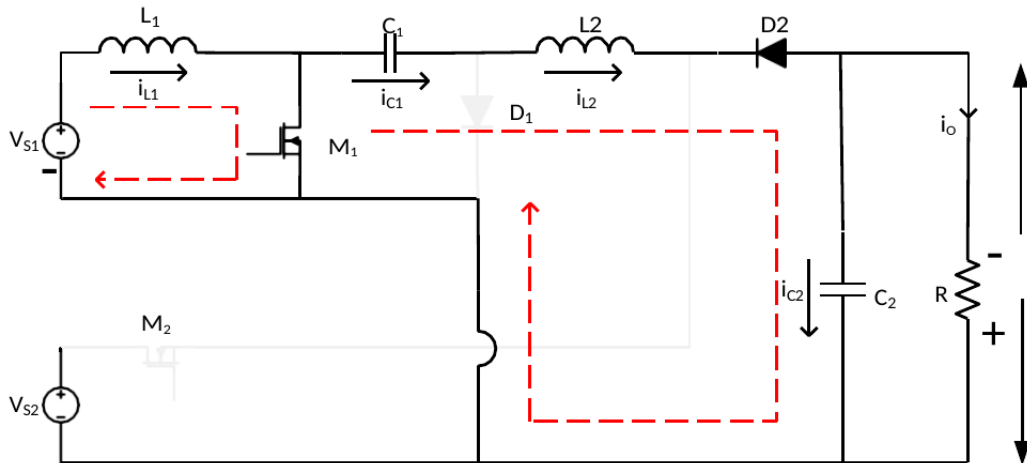
If we represent the states by $x = [i_{L1} \ i_{L2} \ V_{C1} \ V_{C2}]^T$ and the inputs by $u = [V_{S1} \ V_{S2}]^T$, then the dynamics of the mode can be represented by:

$$A_1 = \begin{bmatrix} -\frac{r_{L1}}{L_1} & 0 & 0 & 0 \\ 0 & -\frac{r_{L2} + r_{C1}}{L_2} & -\frac{1}{L_2} & 0 \\ 0 & \frac{1}{C_1} & 0 & 0 \\ 0 & 0 & 0 & -\frac{1}{(R + r_{C2})C_2} \end{bmatrix}, \quad B_1 = \begin{bmatrix} \frac{1}{L_1} & 0 \\ 0 & \frac{1}{L_2} \\ 0 & 0 \\ 0 & 0 \end{bmatrix}$$

2.1.2 Mode 2 ($d_2 T_s < t < d_1 T_s$)

S_1 is on and S_2 is off, so current passes through D_2 and not through D_1 (Fig -3). L_1 continues to be charged by

source 1, while L_2 releases its energy into R . The voltage and current relations for this mode are given in Eqs. (5)-(12).


Fig -3: Circuit configuration and current-flow paths during Mode 2 operation ($d2T_s < t < d1T_s$)

 For L_1 :

$$V_{S1} - r_{L1}i_{L1} - L_1 \frac{di_{L1}}{dt} = 0 \quad (5)$$

$$\frac{di_{L1}}{dt} = -\frac{r_{L1}}{L_1}i_{L1} + \frac{1}{L_1}V_{S1} \quad (6)$$

 Apply it to the L_2 equation:

$$L_2 \frac{di_{L2}}{dt} = -V_{C1} - (r_{L2} + r_{C1})i_{L2} + V_{C2} + r_{C2} \frac{R}{R + r_{C2}} \left[i_{L2} - \frac{1}{R}V_{C2} \right] \quad (9)$$

Simplify:

 For L_2 :

$$L_2 \frac{di_{L2}}{dt} = -V_{C1} - (r_{L2} + r_{C1})i_{L2} + V_{C2} + r_{C2}i_{C2} \quad (7)$$

$$\begin{aligned} \frac{di_{L2}}{dt} = & -\frac{1}{L_2}V_{C1} \\ & + \frac{1}{L_2} \left[\frac{Rr_{C2}}{R + r_{C2}} - (r_{L2} + r_{C1}) \right] i_{L2} \\ & + \frac{R}{L_2(R + r_{C2})} V_{C2} \end{aligned} \quad (10)$$

 C_2 capacitor current equation (KCL):

$$\begin{aligned} i_{L2} &= i_{C2} + i_o \\ i_o &= \frac{V_o}{R} \\ V_o &= V_{C2} - r_{C2}i_{C2} \end{aligned}$$

 For C_1 :

$$\frac{dV_{C1}}{dt} = -\frac{1}{C_1}i_{L2} \quad (11)$$

 For C_2 :

 Calculate the C_2 current:

$$i_{C2} = \frac{R}{R + r_{C2}}i_{L2} - \frac{1}{R + r_{C2}}V_{C2} \quad (8)$$

$$\begin{aligned} \frac{dV_{C2}}{dt} = & \frac{R}{C_2(R + r_{C2})}i_{L2} \\ & - \frac{1}{C_2(R + r_{C2})}V_{C2} \end{aligned} \quad (12)$$

$$A_2 = \begin{bmatrix} -\frac{r_{L1}}{L_1} & 0 & 0 & 0 \\ 0 & \frac{1}{L_2} \left[\frac{Rr_{C2}}{R+r_{C2}} - (r_{L2} + r_{C1}) \right] & -\frac{1}{L_2} & \frac{R}{L_2(R+r_{C2})} \\ 0 & -\frac{1}{C_1} & 0 & 0 \\ 0 & \frac{R}{C_2(R+r_{C2})} & 0 & -\frac{1}{C_2(R+r_{C2})} \end{bmatrix}, \quad B_2 = \begin{bmatrix} \frac{1}{L_1} & 0 \\ 0 & \frac{1}{L_2} \\ 0 & 0 \\ 0 & 0 \end{bmatrix}$$

2.1.3 Mode 3 ($d_1T_s < t < T_s$)

During this interval both switches S_1 and S_2 are open (see Fig -4), and diodes D_1 and D_2 conduct. Only source V_{g2}

supplies the circuit; source V_{g1} is inactive. Energy from source V_{g2} is transferred to L_2 ; L_1 together with the coupling stage releases its stored energy into R .

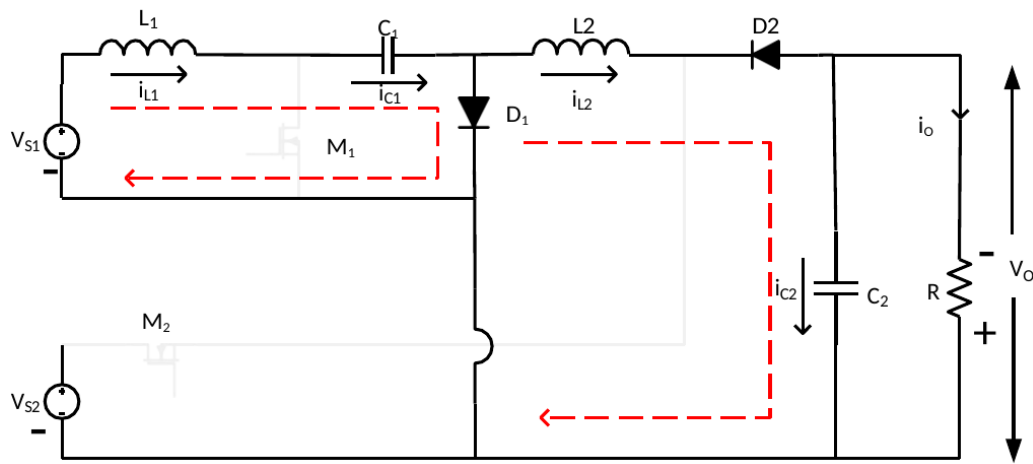


Fig -4: Circuit configuration and current-flow paths during Mode 3 operation ($d_1T_s < t < T_s$)

For L_1 (KVL):

$$V_{S1} - (r_{L1} + r_{C2})i_{L1} - V_{C1} - L_1 \frac{di_{L1}}{dt} = 0 \quad (13)$$

For C_1 (KCL):

$$\frac{dV_{C1}}{dt} = \frac{1}{C_1} i_{L1} \quad (16)$$

$$\frac{di_{L1}}{dt} = -\frac{r_{L1} + r_{C2}}{L_1} i_{L1} - \frac{1}{L_1} V_{C1} + \frac{1}{L_1} V_{S1} \quad (14)$$

For C_2 :

$$\frac{dV_{C2}}{dt} = \frac{R}{C_2(R+r_{C2})} i_{L2} - \frac{1}{C_2(R+r_{C2})} V_{C2} \quad (17)$$

For L_2 (KVL with substitution of currents in capacitors):

$$\frac{di_{L2}}{dt} = -\frac{1}{L_2} \left[r_{L2} + \frac{Rr_{C2}}{R+r_{C2}} \right] i_{L2} - \frac{R}{L_2(R+r_{C2})} V_{C2} \quad (15)$$

$$A_3 = \begin{bmatrix} -\frac{r_{L1} + r_{C2}}{L_1} & 0 & -\frac{1}{L_1} & 0 \\ 0 & -\frac{1}{L_2} \left[r_{L2} + \frac{Rr_{C2}}{R + r_{C2}} \right] & 0 & -\frac{R}{L_2(R + r_{C2})} \\ \frac{1}{C_1} & 0 & 0 & 0 \\ 0 & \frac{R}{C_2(R + r_{C2})} & 0 & -\frac{1}{C_2(R + r_{C2})} \end{bmatrix}, \quad B_3 = \begin{bmatrix} \frac{1}{L_1} & 0 \\ 0 & 0 \\ 0 & 0 \\ 0 & 0 \end{bmatrix}$$

The volt-second balance principle requires the average inductor voltage to be zero over each switching period T_s . Applying it across the three operating modes, the average voltages of L_1 and L_2 give the contribution of each source to the voltage gain in Eqs. (18)–(22).

For L_1 : ON (switch closed):

$$V_{L1} = V_{S1} \quad (18)$$

OFF (switch open) – the output voltage is taken as negative:

$$V_{L1} = V_{S1} - (-V_0) = V_{S1} + V_0 \quad (19)$$

Volt-second balance:

$$\begin{aligned} D_1 V_{S1} + (1 - D_1)(V_{S1} + V_0) &= 0 \\ V_0 &= -\frac{V_{S1}}{1 - D_1} \end{aligned} \quad (20)$$

For L_2 :

$$\begin{aligned} D_2 V_{S2} + (1 - D_2)(V_{S2} + V_0) &= 0 \\ V_0 &= -\frac{V_{S2}}{1 - D_2} \end{aligned} \quad (21)$$

Combined output:

$$V_0 = -\frac{V_{S1}}{1 - D_1} - \frac{V_{S2}}{1 - D_2} \quad (22)$$

2.2 State-Space Averaged Model of the Multi-Input Cuk-Type Converter

A generalized state-space averaged model can be built for multi-input Cuk-type buck-boost converters and used to describe the converter's dynamic behaviour. Since the converter works in three modes within a cycle, a separate set of state-space equations is written for each mode. To represent the system as a whole, these three models are combined by state-space averaging, a standard technique

in power electronics. For this purpose the duty ratios d_1 and d_2 serve as the control variables; d_1 is normally larger than d_2 . The averaged model captures the behaviour of the inductor currents, capacitor voltages and switching action across all three modes. The transfer functions follow from the averaged system matrix, and stability can be examined through its eigenvalues. Since canonical transformations leave eigenvalues, eigenvectors and transfer functions unchanged, the averaged model retains the dynamics and stability properties of the converter.

The state-space equations of the three modes are:

$$\begin{aligned} \dot{x}(t) &= A_1 X(t) + B_1 U(t) \\ y(t) &= E_1 X(t) + F_1 U(t) \end{aligned} \quad (23)$$

$$\begin{aligned} \dot{x}(t) &= A_2 X(t) + B_2 U(t) \\ y(t) &= E_2 X(t) + F_2 U(t) \end{aligned} \quad (24)$$

$$\begin{aligned} \dot{x}(t) &= A_3 X(t) + B_3 U(t) \\ y(t) &= E_3 X(t) + F_3 U(t) \end{aligned} \quad (25)$$

In these equations the system matrices are A_1, A_2, A_3 , the input matrices are B_1, B_2, B_3 , the output matrices are E_1, E_2, E_3 and the feedthrough matrices of the three modes are F_1, F_2 and F_3 . The three sets are then averaged over one switching period. The resulting averaged state-space model is:

$$\begin{aligned} \dot{x}(t) &= A_{\text{avg}} X(t) + B_{\text{avg}} U(t) \\ y(t) &= E_{\text{avg}} X(t) + F_{\text{avg}} U(t) \end{aligned} \quad (26)$$

in which the averaged matrices are:

$$\begin{aligned} A_{\text{avg}} &= d_2 A_1 + (d_1 - d_2) A_2 + (1 - d_1) A_3 \\ B_{\text{avg}} &= d_2 B_1 + (d_1 - d_2) B_2 + (1 - d_1) B_3 \\ E_{\text{avg}} &= d_2 E_1 + (d_1 - d_2) E_2 + (1 - d_1) E_3 \\ F_{\text{avg}} &= d_2 F_1 + (d_1 - d_2) F_2 + (1 - d_1) F_3 \end{aligned} \quad (27)$$

To analyse the interaction between the control loops, small perturbations around the steady-state operating point are considered and the nonlinear terms are

neglected, which leads to the following small-signal linearized model:

$$\hat{\mathbf{x}} = \mathbf{A}\hat{\mathbf{x}} + \mathbf{B}\hat{\mathbf{u}} + [(\mathbf{A}_2 - \mathbf{A}_3)\mathbf{X} + (\mathbf{B}_2 - \mathbf{B}_3)\mathbf{U}]\hat{\mathbf{d}}_1 + [(\mathbf{A}_1 - \mathbf{A}_2)\mathbf{X} + (\mathbf{B}_1 - \mathbf{B}_2)\mathbf{U}]\hat{\mathbf{d}}_2 \quad (28)$$

$$\hat{\mathbf{y}} = \mathbf{E}\hat{\mathbf{x}} + \mathbf{F}\hat{\mathbf{u}} + [(\mathbf{E}_1 - \mathbf{E}_2)\mathbf{X} + (\mathbf{F}_1 - \mathbf{F}_2)\mathbf{U}]\hat{\mathbf{d}}_1 + [(\mathbf{E}_2 - \mathbf{E}_3)\mathbf{X} + (\mathbf{F}_2 - \mathbf{F}_3)\mathbf{U}]\hat{\mathbf{d}}_2 \quad (29)$$

The disturbance vectors that depend on the state are defined as:

$$\begin{aligned} \mathbf{Y}_1 &= (\mathbf{A}_2 - \mathbf{A}_3)\mathbf{X} + (\mathbf{B}_2 - \mathbf{B}_3)\mathbf{U} \\ \mathbf{Y}_2 &= (\mathbf{A}_1 - \mathbf{A}_2)\mathbf{X} + (\mathbf{B}_1 - \mathbf{B}_2)\mathbf{U} \end{aligned} \quad (30)$$

The direct feedthrough terms of the output equation are defined as:

$$\begin{aligned} \Psi_1 &= (\mathbf{E}_1 - \mathbf{E}_2)\mathbf{X} + (\mathbf{F}_1 - \mathbf{F}_2)\mathbf{U} \\ \Psi_2 &= (\mathbf{E}_2 - \mathbf{E}_3)\mathbf{X} + (\mathbf{F}_2 - \mathbf{F}_3)\mathbf{U} \end{aligned} \quad (31)$$

Output-voltage response to the duty ratios. Both duty ratios influence the load voltage:

$$\frac{\hat{v}_o(s)}{\hat{d}_1(s)} = \mathbf{E}(\mathbf{sI} - \mathbf{A})^{-1}\mathbf{Y}_1 + \Psi_1 \quad (32)$$

$$\frac{\hat{v}_o(s)}{\hat{d}_2(s)} = \mathbf{E}(\mathbf{sI} - \mathbf{A})^{-1}\mathbf{Y}_2 + \Psi_2 \quad (33)$$

Inductor-current response. The inductor current also varies with the duty ratios:

$$\frac{\hat{i}_{g2}(s)}{\hat{d}_1(s)} = \mathbf{P}(\mathbf{sI} - \mathbf{A})^{-1}\mathbf{Y}_1 + \Psi_3 \quad (34)$$

$$\frac{\hat{i}_{g2}(s)}{\hat{d}_2(s)} = \mathbf{P}(\mathbf{sI} - \mathbf{A})^{-1}\mathbf{Y}_2 + \Psi_4 \quad (35)$$

Transfer function matrix (TFM). The complete set of transfer functions can be arranged as:

$$\begin{bmatrix} \hat{v}_o \\ \hat{i}_{g2} \end{bmatrix} = \begin{bmatrix} G_{11}(s) & G_{12}(s) \\ G_{21}(s) & G_{22}(s) \end{bmatrix} \begin{bmatrix} \hat{d}_1 \\ \hat{d}_2 \end{bmatrix} \quad (36)$$

This form shows that the system is multivariable, since every element $G_{ij}(s)$ describes the relation between one input and one output, as listed below.

$$G_{11}(s) = \mathbf{E}(\mathbf{sI} - \mathbf{A})^{-1}\mathbf{Y}_1 + \Psi_1 \quad (37)$$

$$G_{12}(s) = \mathbf{E}(\mathbf{sI} - \mathbf{A})^{-1}\mathbf{Y}_2 + \Psi_2 \quad (38)$$

$$G_{12}(s) = \mathbf{E}(\mathbf{sI} - \mathbf{A})^{-1}\mathbf{Y}_2 + \Psi_2 \quad (39)$$

$$G_{22}(s) = \mathbf{P}(\mathbf{sI} - \mathbf{A})^{-1}\mathbf{Y}_4 + \Psi_4 \quad (40)$$

3. DESIGN OF PARAMETERS

The passive components of a multi-input Cuk converter are sized according to ripple criteria. The inductor values are determined from the input voltage, the duty ratio and the permitted current ripple, whereas the capacitors are chosen from charge-balance equations and the permitted voltage ripple [9]. The design equations are:

Inductor design (L_1, L_2) – depending on allowable current ripple and switching time period:

$$L_1 = \frac{V_{in1}D_1T_s}{\Delta I_{L1}} \quad (41)$$

$$L_2 = \frac{V_{in2}D_2T_s}{\Delta I_{L2}} \quad (42)$$

Coupling capacitor (C_1) – depending on allowable voltage ripple:

$$C_1 = \frac{I_{L2}D_2T_s}{\Delta V_{C1}} \quad (43)$$

Output capacitor (C_2) – depending on load current and allowable voltage ripple for stability of output:

$$C_2 = \frac{I_oDT_s}{\Delta V_{C2}} \quad (44)$$

Table -2: Converter design parameters

S. No	Parameter	Value
1	Inductor, L_1	200 mH
2	Inductor, L_2	100 mH
3	Capacitor, C_1	50 mF
4	Capacitor, C_2	100 mF
5	Voltage source, V_{g1}	12 V
6	Voltage source, V_{g2}	24 V
7	Load resistor, R_0	10 Ω
8	Output voltage, V_0	54 V

4. INTERACTION MEASURES

For MIMO control systems, interaction measures are used to examine how the inputs affect the outputs. Both DC-gain-based and Gramian-based interaction measures can be considered here. Frequency-specific measures, namely the effective relative gain array and the effective relative energy array, are also discussed to examine input-output interaction at selected frequencies.

4.1 DC Gain-Based Interaction Measures

DC-gain-based interaction measures assess input-output pairing using the steady-state gain matrix of the system. Because this matrix expresses the static relation between inputs and outputs, the analysis is comparatively simple. The Relative Gain Array (RGA) is the most widely used of these DC-gain-based interaction-measure (DCGIM) techniques [6], [10].

4.1.1 Relative Gain Array

RGA computation requires the DC gain matrix G_0 to be nonsingular. For a 2×2 system, the RGA takes the form:

$$\Lambda = \begin{bmatrix} \lambda & 1-\lambda \\ 1-\lambda & \lambda \end{bmatrix} \quad (45)$$

where λ is the relative gain. The diagonal entries indicate the interaction level of the direct pairings, and the off-diagonal entries indicate the cross-interaction. The RGA is computed from the steady-state gain matrix G_0 as follows:

$$\Lambda = G_0 \circ (G^{-1})^T \quad (46)$$

For our case, the DC gain matrix is calculated using MATLAB and is represented by:

$$G_0 = \begin{bmatrix} -343.6224 & -4.7363 \\ -56.4571 & 265.5601 \end{bmatrix} \quad (47)$$

Therefore, in accordance with the above formula, the RGA matrix will be:

$$\Lambda = \begin{bmatrix} 0.9971 & 0.0029 \\ 0.0029 & 0.9971 \end{bmatrix} \quad (48)$$

Criteria for Relative Gain Array (RGA) Pairing

Case 1: $\lambda = 1$. It is the most favorable value for RGA since it indicates no cross interaction of the loops; thus, such pairings as $u1 \rightarrow y1$ and $u2 \rightarrow y2$ would be perfect.

Case 2: $\lambda = 0$. In case the diagonal element equals zero, only cross pairing would be regarded as possible (namely, $u1 \rightarrow y2$, $u2 \rightarrow y1$).

Case 3: $0 < \lambda < 1$. Elements between zero and one indicate partial interaction, which implies that pairing can still occur but the closed-loop gain will be higher; hence, $\lambda = 0.5$ is not appropriate.

Case 4: $\lambda < 0$. This indicates the presence of interaction such that an increase in one input leads to an opposite change in the output and, as a consequence, an unstable system occurs.

Case 5: $\lambda > 1$. It means high sensitivity and poor conditioning of the system although pairing can theoretically occur; nevertheless, it will make the control more complicated because of interaction.

RGA values for the conditions described above:

Diagonal Dominance: The values of diagonal elements ($\lambda_{11}=0.9971$, $\lambda_{22}=0.9971$) are quite close to unity. Hence, the proper combination is $u1 \rightarrow y1$ and $u2 \rightarrow y2$.

Cross Coupling: The values of off-diagonal elements ($\lambda_{12}=0.0029$, $\lambda_{21}=0.0029$) are very small, thus no cross coupling exists between the loops.

Conditioning: There is no element with a negative or very large value; therefore the system is well-conditioned.

5. RESULTS

5.1 Steady-State Output Verification

Simulating the converter with the designed passive components, supply voltages and duty ratios gives a steady-state output voltage of -52 V, as shown in Fig -5. This agrees with the values expected from the gain equation and the volt-second balance approach, and so supports the analytical results. The output voltage ripple remains below 1%, confirming that the capacitances were chosen suitably.

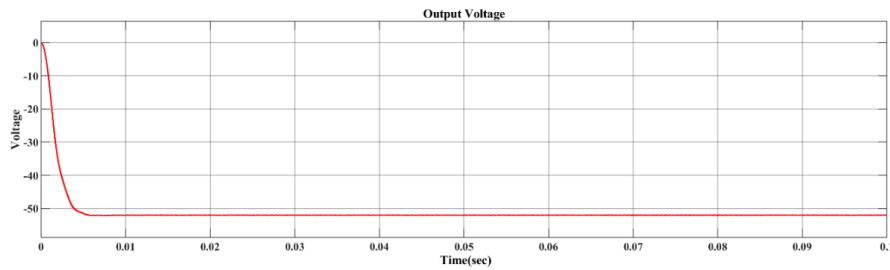


Fig -5: Steady-state response of the converter output voltage.

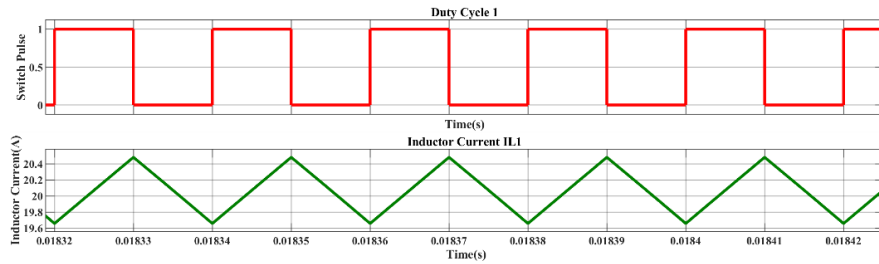


Fig -6: Duty cycle d1 and inductor current iL1 waveforms.



Fig -7: Duty cycle d2 and inductor current iL2 waveforms.

5.2 Inductor Current Behaviour

The inductor currents conduct continuously, as anticipated, and their ripple stays within the 20% value used in the design (Fig -6, Fig -7).

5.3 Dynamic Response to Perturbation

A $\pm 5\%$ perturbation in the duty ratio for d_1 and d_2 caused a proportionate change in the output voltage with little cross-influence. The dynamic transients followed a first-order response whose time constant is consistent with the small-signal model.

5.4 RGA Validation

The validation of the DC gain matrix with respect to the RGA gives:

$$\text{RGA} = \begin{bmatrix} 0.997 & 0.003 \\ 0.002 & 0.998 \end{bmatrix}$$

The diagonal elements (≈ 0.997 and ≈ 0.998) indicate a strong direct input-output relationship, while the off-diagonal elements (≈ 0.002 – 0.003) are very small, confirming weak coupling. A multivariable decoupling controller is therefore unnecessary, since an individual single-loop controller can be used for each duty ratio.

6. CONCLUSION

This paper analysed the interaction among the modelling, design and control of a dual-input Cuk-type buck-boost DC-DC converter for multi-source power conversion. Modifying the Cuk converter to accept two input sources gives a continuous input current, very low output ripple and a regulated inverted output of -52 V. The averaged state-space model was formulated, and the voltage gain was expressed as a function of the duty cycles using the volt-second balance technique. The passive components were selected from the output-ripple requirements. The averaged model was then linearized to obtain the small-signal model, and the RGA analysis showed that the RGA values are close to unity. This allows a simple, independent single-loop control strategy that

eases controller design without sacrificing robustness, so the two inputs are effectively decoupled. The converter is therefore suitable for hybrid power systems with various input sources.

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