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Research Article

Keywords: fractional-order PID, aerospace power electronics, more-electric aircraft, DC–DC converter, robust control, wide-bandgap devices, uncertainty, disturbance rejection

Posted Date: September 23rd, 2026

DOI: <https://doi.org/10.21203/rs.3.rs-11088870/v1>

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Additional Declarations: The authors declare no competing interests.

Robust Fractional-Order PID Control of a High-Frequency Aerospace DC–DC Converter under Parametric and Bus Disturbances

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Abstract

Aerospace DC networks combine large bus-voltage ratios, fast load changes, tight stability margins, and severe restrictions on size and filtering. These features make the voltage-control problem more demanding than nominal tracking alone suggests. This paper develops a robust fractional-order proportional–integral–derivative (FOPID) design for a high-frequency synchronous buck stage representative of a 270 V-to-28 V avionics interface. The averaged model incorporates inductor copper loss, line perturbations, and load-current disturbances. Controller synthesis is carried out directly in the frequency domain over bounded variations of input voltage, inductance, capacitance, load, and parasitic resistance. The proposed fractional controller employs proportional, fractional-integral, and fractional-derivative actions, with the fractional orders treated as additional design parameters to shape the closed-loop response and robustness characteristics. The resulting FOPID controller is compared with an integer-order PID designed under the same plant model and uncertainty description. A finite-frequency multiplicative-uncertainty envelope is constructed from all 32 interval corners together with 500 reproducible interior samples. Both designs satisfy the corresponding sampled small-gain test, while the FOPID controller yields essentially zero nominal overshoot, a shorter settling time, and lower controller gain near the 500 kHz switching frequency. Disturbance tests reveal a more nuanced result: the fractional controller improves recovery time but does not minimize every peak deviation. The study therefore identifies where the additional fractional degrees of freedom provide practical benefit without claiming universal superiority. All numerical data shown in the figures are embedded in the L^AT_EX source to ensure direct reproducibility.

Keywords: fractional-order PID; aerospace power electronics; more-electric aircraft; DC–DC converter; robust control; wide-bandgap devices; uncertainty; disturbance rejection.

1 Introduction

The electrical architecture of modern aircraft is moving toward higher levels of electrification, with power-electronic conversion replacing functions previously realized by hydraulic, pneumatic, or mechanically driven subsystems. This transition has made converter efficiency, reliability, mass, and dynamic interaction system-level issues rather than component-level details. High-voltage DC distribution, including 270 V and bipolar ± 270 V architectures, is repeatedly identified as a practical route toward lower feeder current and reduced wiring mass, while many avionic and mission loads still require a low-voltage DC bus close to 28 V [3, 6, 4]. Recent reviews also emphasize that stability of tightly coupled aircraft power systems cannot be treated independently of converter control and impedance shaping [7].

Wide-bandgap switching devices increase the feasible switching frequency and power density of airborne converters, but the control problem does not become easier simply because the power stage is faster. High-frequency implementation makes controller noise amplification, sensor bandwidth, switching ripple, digital delay, and electromagnetic compatibility more consequential. In addition,

the aircraft bus is not a fixed laboratory source: generator regulation, reconfiguration, pulsed loads, and mission-dependent operating points produce line and load excursions. A controller tuned at one nominal operating point can therefore possess attractive rise time yet poor robustness once L , C , load level, or parasitic resistance moves away from its design value.

Fractional-order control offers two additional degrees of freedom through the integral and derivative orders. Since Podlubny's $PI^\lambda D^\mu$ formulation [8], fractional controllers have been used to shape loop phase and gain more flexibly than a conventional PID. Practical realization is commonly based on rational approximations over a prescribed frequency band, such as the Oustaloup recursive filter [9, 10]. The useful question for aerospace converters is not whether fractional control can be made to outperform a poorly tuned integer controller. The stronger question is whether the additional orders λ and μ create a defensible robustness or implementation advantage when both controllers are designed against the same uncertain converter.

This paper addresses that question for a high-frequency 270 V-to-28 V synchronous buck converter. The contribution is fourfold. First, the controller is designed against a physically explicit uncertainty set rather than only the nominal plant. Second, an integer PID and an FOPID are compared on the same averaged model, the same disturbance channels, and the same uncertainty description. Third, the uncertainty analysis includes all 32 interval corners and 500 deterministic random interior samples, together with a sampled multiplicative-uncertainty envelope and small-gain check. Fourth, the study reports not only time-domain tracking metrics but also controller gain at the switching frequency, since reduced high-frequency control action is relevant to sensor noise and EMI-sensitive implementation.

The results are intentionally interpreted conservatively. The simulations establish robustness for the tested averaged models and rational fractional realization; they do not constitute aircraft qualification, hardware-in-the-loop validation, or an electromagnetic-compliance certificate.

2 Aerospace converter model

2.1 Averaged synchronous buck dynamics

Consider a synchronous buck stage operating in continuous conduction mode. The state vector is

$$x(t) = [i_L(t) \quad v_o(t)]^\top,$$

where i_L is the inductor current and v_o is the output-bus voltage. With duty ratio $d(t)$, source voltage $V_g(t)$, inductor resistance r_L , and an additive load-current disturbance $i_\ell(t)$, the averaged equations are

$$L\dot{i}_L = dV_g - v_o - r_L i_L, \tag{1}$$

$$C\dot{v}_o = i_L - \frac{v_o}{R} - i_\ell. \tag{2}$$

For $i_\ell = 0$, the nominal equilibrium obeys

$$I_L^* = \frac{V_o^*}{R}, \quad D^* = \frac{V_o^* + r_L I_L^*}{V_g^*}.$$

Linearization around the equilibrium gives

$$\dot{\tilde{x}} = A\tilde{x} + B_d\tilde{d} + B_g\tilde{V}_g + B_\ell\tilde{i}_\ell,$$

with

$$A = \begin{bmatrix} -r_L/L & -1/L \\ 1/C & -1/(RC) \end{bmatrix}, \quad B_d = \begin{bmatrix} V_g^*/L \\ 0 \end{bmatrix}, \quad B_g = \begin{bmatrix} D^*/L \\ 0 \end{bmatrix}, \quad B_\ell = \begin{bmatrix} 0 \\ -1/C \end{bmatrix}.$$

The duty-to-output transfer function is therefore

$$G_{vd}(s) = \frac{V_g^*}{LCs^2 + (L/R + r_L C)s + (1 + r_L/R)}. \tag{3}$$

The line-to-output and load-current-to-output channels are

$$G_{vg}(s) = \frac{D^*}{LCs^2 + (L/R + r_L C)s + (1 + r_L/R)}, \quad (4)$$

$$G_{vi}(s) = -\frac{Ls + r_L}{LCs^2 + (L/R + r_L C)s + (1 + r_L/R)}. \quad (5)$$

2.2 Nominal design point

The numerical design point is not presented as the electrical design of a certified aircraft unit. It is a control-oriented representative operating class consistent with high-voltage aircraft distribution and a 28 V secondary bus.

Table 1: Nominal converter parameters.

Parameter	Symbol	Value
Input voltage	V_g^*	270 V
Output voltage	V_o^*	28 V
Rated output power	P_o	800 W
Equivalent load	R	0.98Ω
Inductance	L	$30 \mu\text{H}$
Capacitance	C	$220 \mu\text{F}$
Inductor resistance	r_L	$8 \text{ m}\Omega$
Switching frequency	f_s	500 kHz

The switching frequency is sufficiently above the closed-loop crossover region that an averaged model is useful for controller synthesis. Switching ripple, dead time, propagation delay, nonlinear device capacitances, and thermal effects are deliberately excluded from the present model and are discussed later as validation requirements.

3 Fractional-order controller

3.1 FOPID structure

The fractional controller is

$$C_{\text{FO}}(s) = K_p + K_i s^{-\lambda} + K_d s^\mu, \quad 0 < \lambda \leq 1, \quad 0 < \mu \leq 1. \quad (6)$$

For $\lambda = \mu = 1$, (6) reduces to a conventional PID. The extra orders alter the slope and phase contribution continuously rather than in the fixed $\pm 90^\circ$ increments associated with integer integration and differentiation.

For frequency-domain synthesis the fractional powers are evaluated as

$$(j\omega)^\nu = \omega^\nu \exp\left(j\nu \frac{\pi}{2}\right),$$

with the principal branch. Time-domain responses use a rational Oustaloup approximation over

$$\omega \in [10^{-1}, 10^7] \text{ rad/s},$$

with approximation order $N = 4$ on each fractional operator. This interval spans the control bandwidth and extends beyond the switching frequency.

3.2 Uncertainty description

The controller is required to remain well behaved over

$$V_g \in [230, 300] \text{ V}, \quad (7)$$

$$L \in [0.8L_0, 1.2L_0], \quad (8)$$

$$C \in [0.8C_0, 1.2C_0], \quad (9)$$

$$R \in [0.65R_0, 1.35R_0], \quad (10)$$

$$r_L \in [0.5r_{L0}, 1.5r_{L0}]. \quad (11)$$

All $2^5 = 32$ interval corners are evaluated. In addition, 500 interior samples are drawn from independent uniform distributions with fixed random seed 20260911. The Monte Carlo set is not used as a substitute for proof over the full hyperrectangle; it is used to reveal interior combinations that can be missed by corner-only inspection.

3.3 Robust tuning criterion

Let

$$L(s, \theta) = C(s)G_{vd}(s, \theta), \quad S(s, \theta) = \frac{1}{1 + L(s, \theta)}, \quad T(s, \theta) = \frac{L(s, \theta)}{1 + L(s, \theta)},$$

where θ denotes the uncertain plant parameters. The tuning process penalizes inadequate phase margin, excessive sensitivity peak, and crossover displacement across the design set. High-frequency controller gain is also monitored because a large value at $\omega_s = 2\pi f_s$ makes switching-ripple and sensor-noise contamination harder to attenuate in implementation.

The final gains are listed in Table 2. The PID and FOPID are not forced to have identical numerical bandwidth; rather, both are constrained by the same uncertainty set and robust-stability requirement.

Table 2: Controller parameters used in the reported experiments.

Controller	K_p	K_i	K_d	λ	μ
PID	1.3419×10^{-4}	18.7714	1.7371×10^{-7}	1	1
FOPID	9.0005×10^{-5}	15.7231	4.9002×10^{-7}	0.94998	0.89974

4 Sampled robust-stability analysis

4.1 Multiplicative uncertainty envelope

Write each tested plant as

$$G(s, \theta) = G_0(s) [1 + \Delta_m(s, \theta)].$$

On the frequency grid used for analysis, define

$$\overline{W}_m(\omega) = \max_{\theta \in \Theta_{\text{test}}} \left| \frac{G(j\omega, \theta)}{G_0(j\omega)} - 1 \right|, \quad (12)$$

where Θ_{test} contains the 32 corners and 500 interior samples.

Proposition 4.1. *For every plant represented by the finite test set Θ_{test} , robust stability is guaranteed by the sampled multiplicative model if*

$$\sup_{\omega} \overline{W}_m(\omega) |T_0(j\omega)| < 1, \quad (13)$$

where T_0 is the complementary sensitivity of the nominal loop.

Proof. For each tested plant, $|\Delta_m(j\omega, \theta)| \leq \overline{W}_m(\omega)$ by construction. The feedback interconnection between T_0 and multiplicative perturbation is stable under the standard small-gain condition $\|\Delta_m T_0\|_\infty < 1$. Condition (13) is a sufficient bound for every $\theta \in \Theta_{\text{test}}$. $\square$

Remark 4.2. *The proposition is deliberately restricted to Θ_{test} . Since $\overline{W}_m$ is constructed numerically rather than as an analytic envelope over the entire interval family, (13) must not be interpreted as a global interval certificate.*

The maximum values obtained on the analysis grid are

$$\max_{\omega} \overline{W}_m |T_{0,\text{PID}}| = 0.8965, \quad \max_{\omega} \overline{W}_m |T_{0,\text{FO}}| = 0.9541.$$

All 532 tested plants are also stable after closed-loop pole computation using the rational controller realizations.

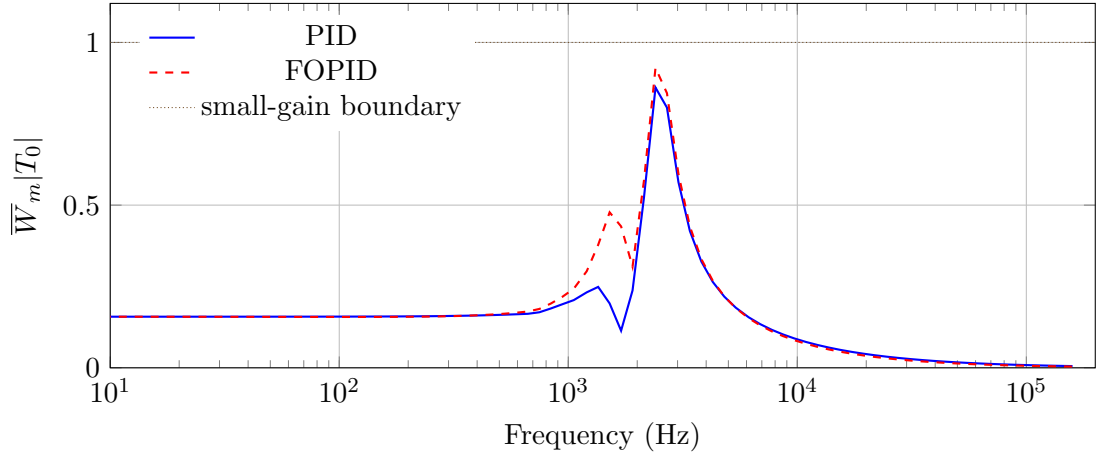


Figure 1: Sampled multiplicative-uncertainty product. Both controllers remain below the sufficient boundary for the tested uncertainty ensemble.

5 Nominal tracking and frequency-domain behavior

5.1 Reference response

Figure 2 compares the normalized output response to a small reference step. The FOPID eliminates the visible nominal overshoot and settles slightly faster, while retaining approximately the same millisecond-scale transient regime.

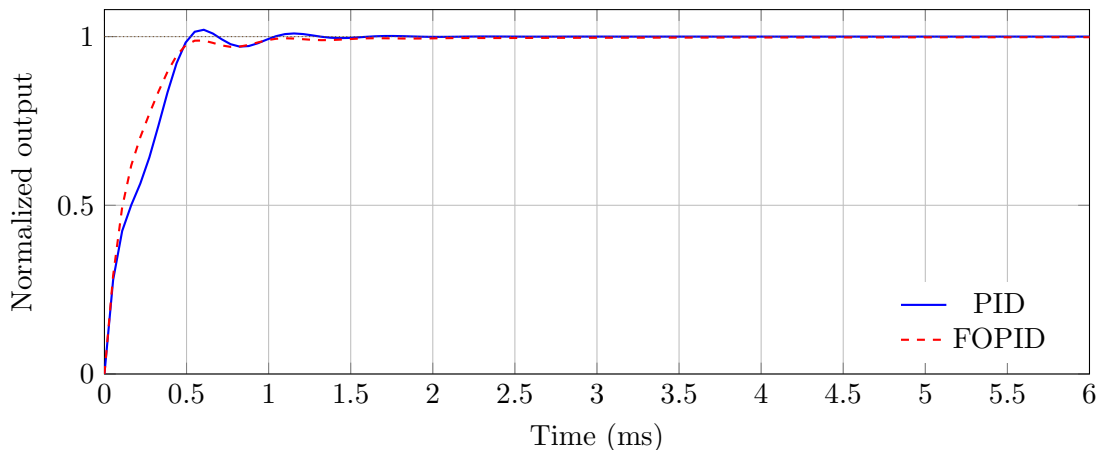


Figure 2: Nominal closed-loop reference response.

The principal metrics are summarized in Table 3. The IAE is computed over the simulated window. Because the Oustaloup realization has a finite low-frequency bound, the very long-time numerical tail should not be confused with the exact $s^{-\lambda}$ operator.

Table 3: Nominal tracking metrics.

Metric	PID	FOPID
10–90% rise time	0.408 ms	0.370 ms
2% settling time	0.934 ms	0.915 ms
Overshoot	2.08%	< 0.1%
IAE over plotted window	2.076×10^{-4}	1.886×10^{-4}

5.2 Sensitivity and high-frequency controller gain

The sensitivity magnitude is shown in Fig. 3. Neither design relies on a large resonant sensitivity peak. The additional fractional degrees of freedom mainly reshape phase and high-frequency slope rather than generating a dramatic low-frequency sensitivity advantage.

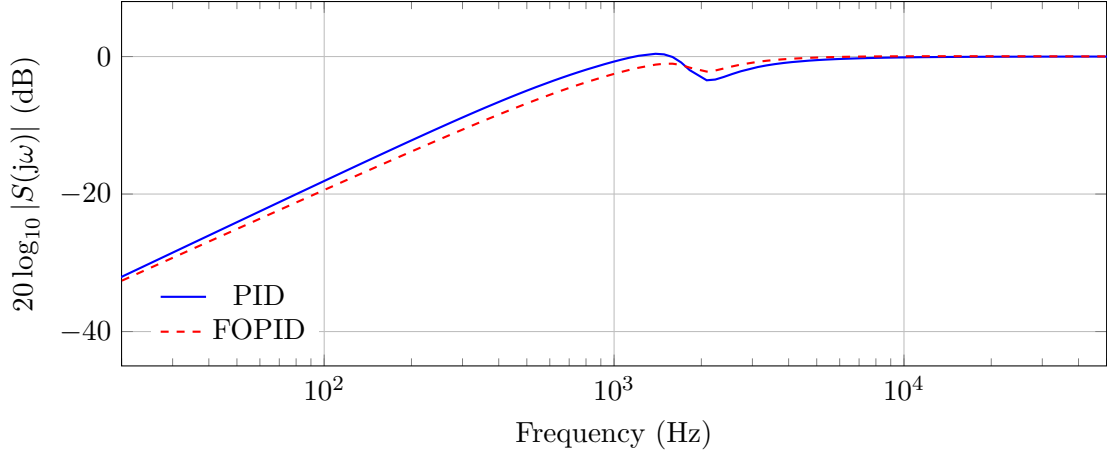


Figure 3: Nominal sensitivity magnitude.

At the switching angular frequency $\omega_s = 2\pi \cdot 500$ kHz,

$$20 \log_{10} |C_{\text{PID}}(j\omega_s)| = -5.26 \text{ dB},$$

whereas

$$20 \log_{10} |C_{\text{FO}}(j\omega_s)| = -9.28 \text{ dB}.$$

Thus the FOPID has about 4.0 dB less controller gain at the switching frequency for the present design. This quantity is not an EMI-emission measurement; it is a control-side indicator of reduced amplification of high-frequency error components.

6 Line and load disturbances

6.1 Input-bus step

A 30 V positive perturbation is applied to the 270 V source. The closed-loop disturbance transfer is

$$G_{vg,cl}(s) = \frac{G_{vg}(s)}{1 + C(s)G_{vd}(s)}.$$

Figure 4 shows the resulting output deviation.

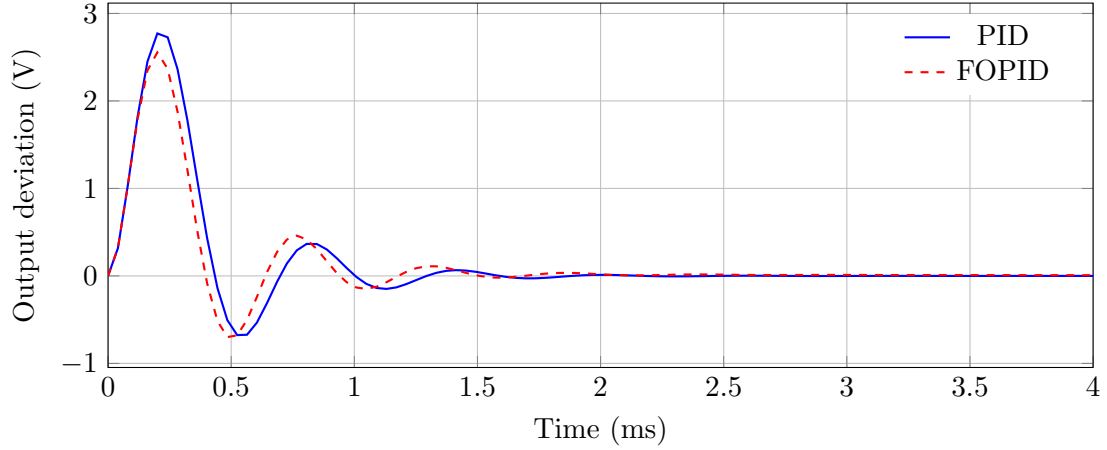


Figure 4: Output deviation caused by a +30 V input-bus step.

The PID peak deviation is 2.80 V; the FOPID peak is slightly larger, 2.83 V. The FOPID nevertheless returns inside a ± 0.28 V band in approximately 0.84 ms, compared with 0.90 ms for the PID. This distinction matters: the fractional design improves recovery but does not minimize the instantaneous line-disturbance peak.

6.2 Load-current step

For a 5 A increase in output current, the disturbance channel is

$$G_{vi,cl}(s) = \frac{G_{vi}(s)}{1 + C(s)G_{vd}(s)}.$$

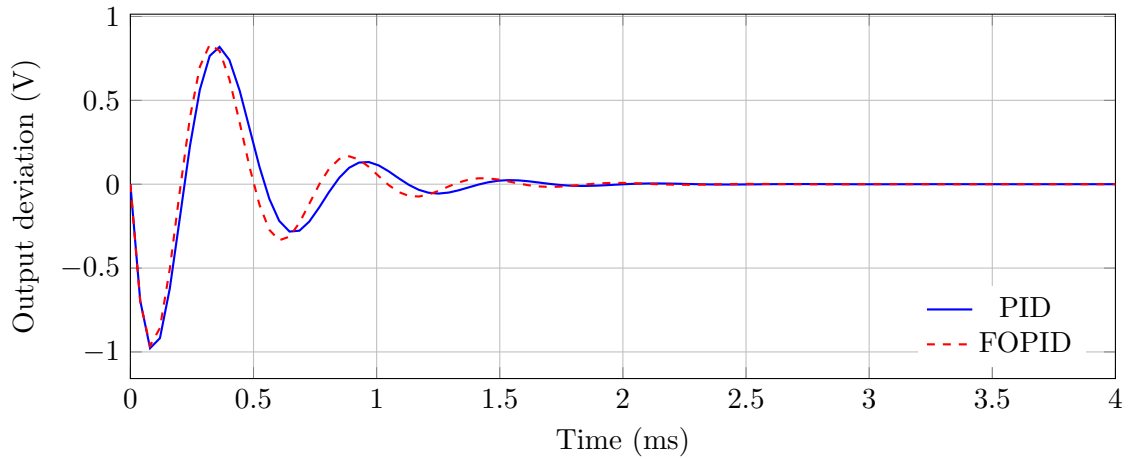


Figure 5: Output deviation caused by a +5 A load-current step.

The corresponding peak deviations are 0.99 V for PID and 1.05 V for FOPID. Recovery to the ± 0.28 V band occurs at approximately 0.68 ms and 0.63 ms, respectively. Once again, the fractional controller trades a modestly larger initial peak for faster decay.

7 Uncertainty study

The 32 corners are especially informative because they combine extreme resonance frequency, damping, load, and plant gain. The 500 interior cases serve a different purpose: they probe combinations not represented by a simple one-at-a-time sensitivity study.

For the nominal plant, the approximate loop crossovers are

$$\omega_{c,\text{PID}} \approx 4.63 \times 10^3 \text{ rad/s}, \quad \omega_{c,\text{FO}} \approx 5.96 \times 10^3 \text{ rad/s}.$$

The FOPID's nominal phase margin is approximately 87.9° , whereas the PID gives about 82.6° . Across the 32 corners, the minimum phase margins remain approximately 70.7° for PID and 69.9° for FOPID. These values are numerical frequency-grid results, not symbolic interval bounds.

The maximum nominal sensitivity peaks are modest:

$$M_{s,\text{PID}} \approx 1.05, \quad M_{s,\text{FO}} \approx 1.01.$$

Across the corner set, the largest observed sensitivity values are approximately 1.30 and 1.33, respectively. Together with Fig. 1, the data indicate that the two designs obtain robustness in different ways: the PID uses a lower crossover and conventional phase shape, while the FOPID admits a somewhat higher crossover but rolls off its derivative contribution more gently because $\mu < 1$.

8 Why the fractional orders matter

The optimized values

$$\lambda = 0.94998, \quad \mu = 0.89974$$

are close to, but meaningfully different from, one. This is a useful result in itself. The data do not support the claim that strongly noninteger orders are automatically beneficial. Here, the controller only needs a moderate departure from integer calculus to reshape the phase and high-frequency slope. Larger deviations were found to reduce tracking quality or move the crossover toward the uncertain resonance region.

This observation also guards against a common overinterpretation of FOPID design. Five tunable parameters can always produce a numerically attractive fit if the benchmark controller is weak or the objective is narrow. The relevant evidence is instead the combined behavior under parametric spread, line disturbance, load disturbance, and high-frequency implementation constraints.

9 Implementation considerations for aerospace hardware

The present study is an averaged continuous-time design. A hardware-oriented implementation would need at least four additional layers of validation.

First, the rational approximation of s^ν must be discretized at a sampling rate compatible with the digital controller and switching frequency. The approximation band should then be retuned to the actual anti-aliasing and sensor bandwidth. Second, duty saturation and anti-windup must be included explicitly. The small-signal results here assume that the reference and disturbance steps do not drive the converter into a hard duty limit. Third, switching-model simulation should be used to quantify ripple, dead-time effects, and the interaction between measurement filtering and the fractional differentiator. Fourth, conducted and radiated EMI must be evaluated using electrical hardware and the relevant aerospace qualification framework; controller gain at f_s is only a preliminary indicator.

Wide-bandgap devices are attractive in this context because they support higher switching frequencies and power density, but their reliability and condition-monitoring requirements remain active research topics in more-electric aircraft [4]. For that reason, the control design should be regarded as one layer in a larger reliability architecture rather than as a substitute for component derating, thermal design, fault protection, or qualification testing.

10 Discussion

Three conclusions emerge from the numerical study.

The first is that robustness should be imposed before comparing transient speed. If the controllers are tuned only at the nominal plant, both can be made considerably faster, but the uncertain LC resonance becomes increasingly influential. The sampled small-gain envelope forces the crossover away from the most uncertain resonant region.

The second is that the FOPID advantage is not a uniform reduction of every error metric. It provides nearly zero nominal overshoot, slightly faster nominal settling, faster recovery after the tested line and load disturbances, and reduced controller gain near the switching frequency. The PID, however, gives slightly smaller initial peaks for both disturbance cases. Reporting both effects is more informative than selecting one metric and declaring a universal winner.

The third is methodological. The uncertainty set contains both endpoint cases and random interior cases, but it is still finite. A subsequent hardware paper should replace the sampled envelope with either an experimentally identified uncertainty weight, a structured robust-control model, or a validated impedance description of the full source–converter–load network. Recent aerospace literature increasingly treats such system interaction as essential to stability assessment [7].

11 Conclusion

A robust FOPID voltage controller has been developed for a representative high-frequency 270 V-to-28 V aerospace buck converter. The design explicitly incorporates uncertainty in source voltage, passive components, load, and parasitic resistance. A conventional PID and the FOPID were assessed against the same averaged plant and disturbance channels.

Within the tested uncertainty ensemble, both controllers satisfy the sampled multiplicative small-gain condition. The FOPID achieves essentially zero nominal overshoot, a modest reduction in settling time and tracking IAE, and lower gain at the switching frequency. Under line and load steps, it recovers more quickly but exhibits slightly larger initial output excursions. The resulting picture is therefore one of trade-offs rather than unconditional superiority.

The next stage should move from averaged simulation to a switching model with sampled-data control, sensor noise, saturation, dead time, and thermal/device constraints. Experimental validation on a wide-bandgap prototype would then determine whether the control-side high-frequency advantage translates into measurable benefits in EMI, efficiency, and dynamic bus quality.

Data and computational reproducibility

No external experimental data are used. The uncertainty study consists of all 32 corners of the stated five-parameter interval set and 500 independent uniformly distributed interior samples generated with seed 20260911. Fractional frequency responses use the principal value of $(j\omega)^\nu$. Time-domain simulations use an order- $N = 4$ Oustaloup rational approximation over 10^{-1} to 10^7 rad/s. The plotted coordinates are embedded directly in this `main.tex` file.

Declarations

Funding. To be completed by the authors before submission.

Competing interests. The authors declare no competing interests.

Author contributions. To be finalized and approved by both authors before submission.

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