

Article

Parametric Study of an H-Shaped-Core Magnetic Field Energy Harvester for Railway Traction-Returning Magnetic Fields

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Abstract

During train operation, railway traction-returning current generates a power-frequency magnetic field around the rail, offering a potential energy source for self-powered track-side monitoring nodes. The H-shaped-core magnetic field energy harvester (MFEH) is attractive because it can be installed beneath the rail without enclosing the conductor, yet its output is strongly affected by the coupled rail-core-coil system. To clarify these effects, a three-dimensional electromagnetic-circuit-coupled finite-element model of an experimentally validated laminated-silicon-steel H-shaped-core MFEH was established to examine core and coil parameters. Increasing the center-leg and side-leg lengths weakens demagnetization but intensifies eddy-current losses, causing output power to approach saturation. Under a 50 Hz, 300 A current in a 54E1 rail and series-tuned matching, output power approaches 5.1 W beyond a center-leg length of 1000 mm and 3.25 W beyond a side-leg length of 700 mm. Within the investigated ranges, center-leg and side-leg lengths of approximately 800 and 400 mm provide the best power–volume performance, respectively. Increasing side-leg height or width also improves output. A larger coil span improves output by reducing internal resistance, whereas more turns yield diminishing gains because of higher winding and eddy-current losses. These findings provide a quantitative basis for parametric design of H-shaped-core MFEHs in railway environments.

Keywords: energy harvesting; magnetic field energy harvester; traction-returning magnetic field; self-powered

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1. Introduction

With the rapid development of intelligent railway systems, long-term and continuous condition monitoring of critical infrastructures, including rails, subgrades, bridges, tunnels, and other track-supporting structures, has become an important means of ensuring operational safety and improving maintenance efficiency [1–3]. Distributed wireless sensor networks have shown strong potential for railway health monitoring because of their flexible deployment, high spatial density, and strong adaptability [4–6]. However, as the number of monitoring nodes continues to increase, the issue of long-term and reliable power supply becomes increasingly pressing. Wired power supply requires long-distance cabling, leading to high installation and maintenance costs and considerable interference with existing railway operations. Battery-powered solutions, although more

flexible, are constrained by limited energy capacity and service life and therefore require frequent maintenance [4–6]. Accordingly, harvesting ambient energy available in the railway environment to achieve self-powered monitoring nodes has become an important route toward the long-term autonomous operation of railway monitoring systems [7].

In recent years, energy harvesting in railway environments has attracted extensive attention, with existing studies mainly focusing on solar, wind, thermal, and vibration energy [7–9]. However, the output of these conventional energy harvesting approaches is usually affected by external conditions such as solar irradiance [10,11], wind speed [12,13], temperature gradient [14,15], and vibration level [16–18], which limits their continuity and stability in complex railway service environments. There is therefore a clear need to explore a new harvesting approach based on an energy source that is more stable, less dependent on ambient weather conditions, and better suited to the harsh railway environment. Representative railway energy-harvesting technologies are compared in Table 1 in terms of their configurations, operating conditions, and reported outputs.

In alternating-current (AC) electrified railways, the traction current enters the train through the pantograph, returns through the rail track, and generates a stable power-frequency alternating magnetic field around the rail. This magnetic field, generated by the traction-returning current, is directly associated with train operating conditions and is largely unaffected by weather or day–night variations, making it a promising non-contact energy source for low-power trackside monitoring nodes [19,20]. In practical railway applications, the traction-returning magnetic field is intermittent because of the discontinuous passage of trains; therefore, the proposed magnetic field energy harvester (MFEH) is more suitable for energy-buffered, low-duty-cycle trackside monitoring nodes. Magnetic-field energy can be converted into electricity through magneto-mechano-electric coupling [21] or, as considered in the present study, directly through electromagnetic induction using an MFEH composed of a high-permeability magnetic core and an induction coil. Existing studies have demonstrated the feasibility of this technical route in railway applications.

From the perspective of installation configuration, MFEHs reported for railway applications can be broadly grouped into three categories. The first category comprises rail-bottom-mounted harvesters. Kuang et al. [19] proposed a ferrite-core harvester consisting of a central magnetic rod and two square flux-collecting end plates. The end plates partially surrounded the rail so as to gather the high-intensity magnetic field around the rail foot, and their study showed that the optimal design should balance effective permeability against eddy-current loss. To further improve mechanical reliability and installation adaptability, Kuang et al. [22] subsequently developed an H-shaped laminated silicon-steel core formed by a central leg and two side legs with identical cross-sections. This design eliminated the rail-enclosing configuration and was directly clamped beneath the rail, with its performance verified through field tests. The second category comprises trackside free-standing harvesters. Espe et al. [23] developed a dumbbell-shaped magnetic core consisting of two steel discs and ferrite rods positioned between them, which enabled non-intrusive energy harvesting when placed freely at the trackside, approximately 0.5 m away from the rail. The third category comprises onboard-mounted harvesters. Tao et al. [24], for freight-train sensor applications, proposed mounting the harvester beneath the bogie and systematically investigated the effects of compensation topology and magnetic-core parameters. Among these configurations, the H-shaped open-magnetic-circuit core is particularly attractive for railway deployment because it does not surround the rail, has a compact structure, and can be installed independently beneath the rail.

From a methodological point of view, the analysis of MFEHs must simultaneously account for magnetic flux concentration in the core, induced voltage in the coil, eddy-current loss in the rail, and the response of the external circuit [25]. Traditional analytical models are generally insufficient to capture the spatial magnetic-field distribution and

terminal characteristics of such systems in a complete manner. In comparison, the electromagnetic-circuit coupled finite element method (FEM) is capable of handling the multiphysics coupling in the rail-core-coil-circuit system more effectively. In particular, the coupled finite element model established by Kuang et al. [19,25] can simultaneously predict internal resistance, internal inductance, open-circuit voltage, and output power and has become an important approach for analyzing and optimizing such harvesters.

Collectively, previous studies have established the feasibility of railway magnetic-field energy harvesting and provided an electromagnetic-circuit-coupled modelling basis for analyzing the interactions among the rail, magnetic core, coil, and external circuit. In particular, the experimentally validated H-shaped-core MFEH reported by Kuang et al. [22] provides a suitable platform for further parametric investigation. However, a systematic understanding of how individual core and coil parameters govern magnetic-flux concentration, demagnetization, winding resistance, eddy-current loss, and impedance-matched output is still lacking. In particular, the transitions from rapid power growth to diminishing returns, together with the associated power–volume trade-offs, have not been quantitatively characterized over sufficiently broad parameter ranges. A systematic parametric investigation is therefore required to complement the validated prototype and provide broader guidance for the design of H-shaped-core railway MFEHs.

To address this gap, the present study establishes a quantitative basis for parametric design using a three-dimensional electromagnetic-circuit-coupled finite-element model. The effects of center-leg length, side-leg length, and cross-sectional dimensions, coil span, and number of coil turns are investigated through controlled one-factor-at-a-time analyses. The selected parameter ranges are sufficiently broad to capture, where applicable, the transition from rapid performance growth to diminishing returns. The analysis separately evaluates winding resistance and equivalent eddy-current resistance, distinguishes eddy-current losses in the rail and magnetic core, and relates geometry-dependent changes in effective permeability to the demagnetizing effect. The core–coil assembly volume and volumetric output power density are further introduced to distinguish absolute-power saturation from favorable power–volume design points within the investigated ranges. Through these analyses, this study clarifies the mechanisms governing diminishing power gains and provides quantitative guidance for the structural and coil design of railway MFEHs.

Table 1. Comparison of representative railway energy-harvesting technologies.

Energy Source	Configuration	Condition	Reported Output	Ref.
Traction-returning magnetic field	H-shaped laminated core; rail-bottom	100–500 A (lab); train passage (field)	0.184–5.178 W (lab); 1.138 ± 1.603 J/train (field)	[22]
Traction-returning magnetic field	Free-standing magnetic core; trackside	200 A, 50 Hz (lab); freight-train passage (field)	40.5 mW (lab); 109 mJ/train (field)	[23]
Rail vibration	Rail-borne piezoelectric harvester	Random rail vibration	1.037 mW Hz ^{−1} (first-mode PSD peak)	[17]
Rail–ground temperature difference	TEG beneath the rail	ΔT = 8–29.2 °C	5.8 mW (field); 316.8 mW (lab)	[14]
Solar irradiation	Foldable photovoltaic system; trackside	Railway-site assessment	Up to 16.7 kWh year ^{−1} (estimated)	[10]

Note: PSD denotes power spectral density. TEG denotes thermoelectric generator. The reported values were obtained under different operating conditions and evaluation metrics and should not be interpreted as a direct performance ranking.

2. Railway Traction-Return Magnetic Field and H-Shaped-Core MFEH

2.1. Railway Traction-Returning Current and Its Magnetic Field

Electrified railways, featuring high transport capacity, high efficiency, reliable operation, and low environmental impact, have become a major form of modern rail transportation [26]. Figure 1a shows the 25 kV single-phase AC traction power supply system widely adopted in electrified railways worldwide. In this system, the traction current is delivered from the traction substation to the locomotive through the overhead contact system. After passing through the train, the current enters the rail via the wheel–rail interface and returns to the traction substation through the rail and the associated return-current paths. Part of the current may return through dedicated return conductors, while another part may be diverted through the earth because of rail–earth coupling [27]. Under AC operation, the traction-returning current typically ranges from several hundred amperes to more than one thousand amperes, generating a stable power-frequency alternating magnetic field in the vicinity of the rail, as shown in Figure 1b. Owing to the skin effect, the current distribution within the rail is non-uniform, with relatively high current density near the rail-head edges and rail-foot corners, where the surrounding alternating magnetic field is correspondingly stronger. Through electromagnetic induction, this magnetic field energy can be converted into electrical energy to power wireless sensing devices in railway monitoring systems.

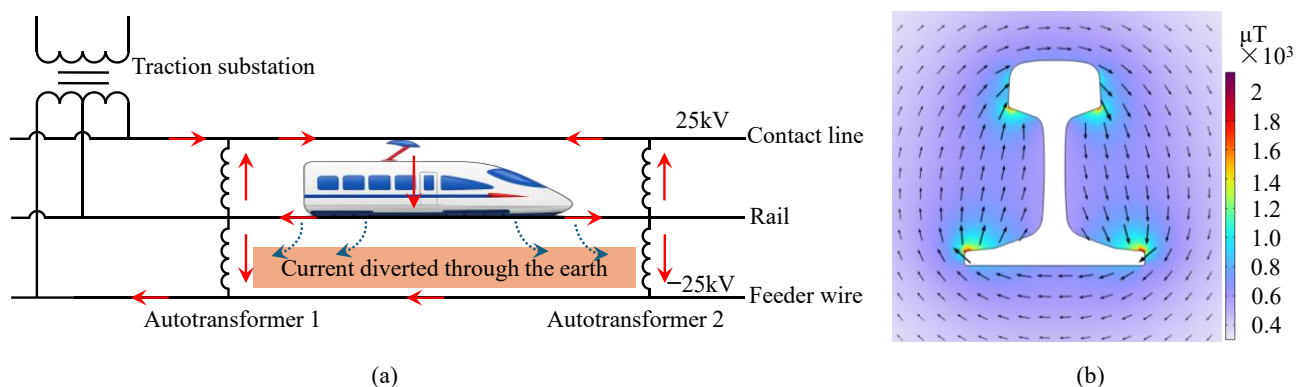


Figure 1. Railway traction-returning magnetic field: (a) schematic of the 25 kV 50/60 Hz alternating-current railway electrification system; (b) magnetic flux density around the rail under a 50 Hz, 300 A traction-returning current. The arrows in (b) indicate the direction of the magnetic flux density.

2.2. MFEH Based on an H-Shaped Core

The MFEH studied in this work consists of an H-shaped core and a coil, as shown in Figure 2a. The core is composed of a center leg and two side legs, forming an open magnetic circuit. The coil is wound around the center leg to couple the time-varying magnetic flux in this region and convert it into an open-circuit-induced voltage, thereby enabling energy harvesting from the alternating magnetic field in railway environments.

Figure 2b illustrates the magnetic flux distribution in the MFEH. Under the alternating magnetic field generated by the traction-returning current, the side legs collect and guide magnetic flux toward the center-leg region, which forms the main flux path of the MFEH. The coil encloses this main flux path and links the time-varying magnetic flux, thereby generating an open-circuit voltage. The magnetic flux then extends from the center leg toward both ends, splits through the side legs, and finally closes through the surrounding air domain. Since the core adopts an open-magnetic-circuit configuration and cannot fully enclose the current-carrying conductor, demagnetization occurs and reduces the effective permeability [22,28]. In addition, the magnetic flux distribution near the ends

and corners is generally non-uniform and is accompanied by a certain degree of flux leakage [29].

Based on the above structural characteristics and flux-transfer mechanism, the key factors governing the H-shaped-core-coil system can be grouped into core geometric parameters and coil parameters. The core parameters include the center-leg dimensions $l_c \times w_c \times h_c$, and the side-leg dimensions $l_s \times w_s \times h_s$, whereas the coil parameters include the number of turns N and the axial coil span. The definitions of these geometric parameters are illustrated in Figure 2a, and the parameters varied in the subsequent analyses are specified in the corresponding sections. These parameters are taken as independent variables in the subsequent sections to systematically assess their effects on the electromagnetic response and output performance of the MFEH.

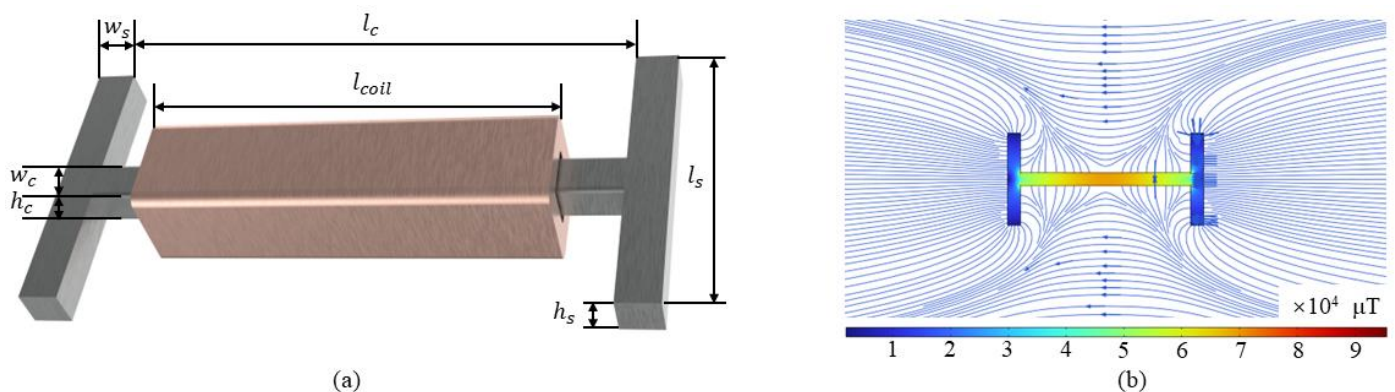


Figure 2. H-shaped-core magnetic field energy harvester (MFEH) and its magnetic flux distribution: (a) prototype of the MFEH; (b) magnetic flux distribution in the MFEH.

2.3. Fundamental Principle of Magnetic Field Energy Harvesting

When the traction-returning current $i(t) = I_r \sin(\omega t)$ flows through the rail, an alternating external magnetic field H_{ex} is established at the MFEH location. When the MFEH is connected to an external load, its terminal behavior can be represented by an equivalent circuit comprising an open-circuit voltage U_{oc} , an internal inductance L , and an internal resistance R , which correspond to the coil inductance and coil resistance in the terminal equivalent circuit, respectively, as shown in Figure 3 [25].

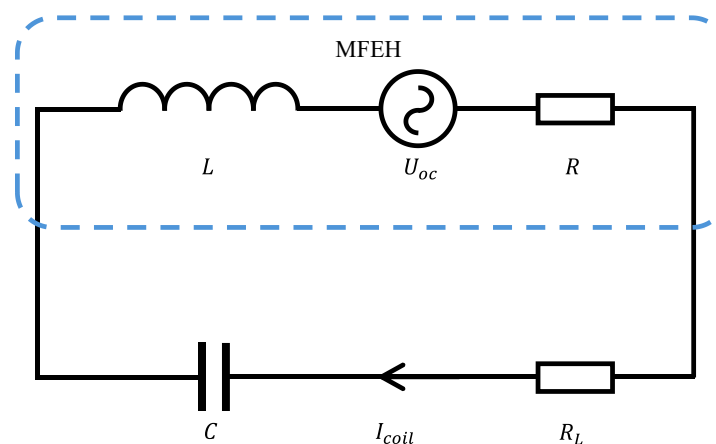


Figure 3. Equivalent circuit model of the magnetic field energy harvester (MFEH). The dashed box denotes the equivalent electrical model of the MFEH.

Under the ideal open-circuit condition, the coil terminals are disconnected from the external load, and the coil current is zero. This treatment is consistent with previous mag-

netic-field energy harvesting studies [19]. Therefore, the magnetic field produced by the coil current is neglected in the calculation of the open-circuit flux density and open-circuit voltage. The amplitude of the open-circuit voltage is then given by [28]:

$$U_{oc} = N\omega A_e B_{oc} = N\omega A_e \mu_{e,r} B_{ex} \quad (1)$$

where N is the number of coil turns, $\omega = 2\pi f$ is the angular frequency, A_e is the effective coil cross-sectional area, and $\mu_{e,r}$ is the effective relative permeability of the magnetic core, hereafter referred to as the effective permeability. B_{oc} denotes the open-circuit flux density, defined as the average magnetic flux density over the middle cross-section of the center leg when the MFEH is subjected to the external alternating magnetic field with the coil open-circuited. B_{ex} denotes the external flux density, usually taken as the average magnetic flux density at the same location under the same excitation after removing the magnetic core. Therefore,

$$B_{oc} = \mu_{e,r} B_{ex} = \mu_{e,r} \mu_0 H_{ex} \quad (2)$$

where μ_0 is the permeability of free space. The effective relative permeability can be approximately related to the intrinsic relative permeability μ_r by [28],

$$\mu_{e,r} \approx \frac{\mu_r}{1 + N_d(\mu_r - 1)} \quad (3)$$

where N_d is the equivalent demagnetization factor, reflecting the constraint of demagnetization and flux leakage on the magnetization response in an open magnetic circuit. In the parametric analyses, $\mu_{e,r}$ was calculated from the FEM-derived B_{oc} and B_{ex} using Equation (2), and N_d was subsequently obtained from Equation (3).

Under the linear approximation, the internal inductance L of the MFEH can be approximately expressed as [30]

$$L = \frac{\mu_0 \mu_{e,r} N^2 A_e}{l_{coil}} \quad (4)$$

where l_{coil} is the coil span. Equation (4) is used to describe the approximate dependence of L on the effective permeability, number of turns, effective area, and coil span. In the subsequent parametric analyses, L is extracted from the finite-element terminal voltage–current relationship. The internal impedance of the MFEH is

$$Z_{EH} = R + j\omega L \quad (5)$$

where R is the internal resistance of the MFEH, including the copper loss of the winding R_{wire} and the magnetic-field-related loss resistance R_{loss} . In the finite-element model, eddy-current effects are included by assigning electrical conductivity to the conductive domains, including the rail and the magnetic core [19,25]. Therefore, the eddy-current loss considered in this work includes both the rail eddy-current loss and the magnetic-core eddy-current loss. In general, magnetic-core losses may also include hysteresis loss and excess loss. Following previous low-frequency MFEH modelling studies, these non-eddy-current-core-loss components are neglected under the 50 Hz operating condition considered here [28,31]. Accordingly, R_{loss} is represented by the equivalent eddy-current resistance R_{eddy} , which can be written as [25]

$$R_{eddy} = \frac{2P_{eddy}}{I_{coil}^2} \quad (6)$$

where $P_{eddy} = P_{r,e} + P_{c,e}$ is the total eddy-current loss power, with $P_{r,e}$ and $P_{c,e}$ denoting the eddy-current losses in the rail and the magnetic core, respectively, and I_{coil} is the coil-current amplitude. Under coil-current excitation, the alternating magnetic field induces eddy currents in nearby conductive magnetic components, including the adjacent

rail and the magnetic core. The reverse magnetic field generated by these eddy currents reduces the flux variation linked with the coil and appears at the circuit level as an additional equivalent resistance [25]. Therefore, the internal resistance of the MFEH can be approximated as

$$R \approx R_{\text{wire}} + R_{\text{eddy}} = \frac{\rho l_{\text{wire}}}{S_{\text{wire}}} + \frac{2P_{\text{eddy}}}{I_{\text{coil}}^2} \quad (7)$$

where ρ is the electrical resistivity of the wire, l_{wire} is the total wire length, and S_{wire} is the wire cross-sectional area.

When a pure resistive load R_L is connected, a tuning capacitor C is usually connected in series to form a series resistor–inductor–capacitor (RLC) circuit so as to compensate for the inductive reactance and maximize the output power. The total impedance of the circuit Z_{tot} is

$$Z_{\text{tot}} = (R + R_L) + j(\omega L - \frac{1}{\omega C}) \quad (8)$$

At the target frequency, the condition for series resonance is

$$\omega L - \frac{1}{\omega C} = 0 \Rightarrow C = \frac{1}{\omega^2 L} \quad (9)$$

When the tuning condition is satisfied and $R_L = R$, the maximum output power is obtained as

$$P_{\text{max}} = \frac{U_{\text{oc}}^2}{8R} = \frac{N^2 \omega^2 A_e^2 \mu_0^2 \mu_{e,r}^2 H_{\text{ex}}^2}{8R} \quad (10)$$

3. Parametric Design Methodology Based on Finite Element Modelling

3.1. Basic Setup of the Finite Element Model

To investigate the effects of key design parameters on the MFEH, a three-dimensional electromagnetic-circuit-coupled finite-element model of the rail-MFEH-air-domain system was established in COMSOL Multiphysics® version 6.2, as shown in Figure 4b. The model was evaluated under different electromagnetic excitation and external-circuit conditions to obtain the magnetic-field distribution, equivalent-circuit parameters, open-circuit response, and impedance-matched output power. The specific excitation and circuit conditions used for the different performance indicators are described in Section 3.2.

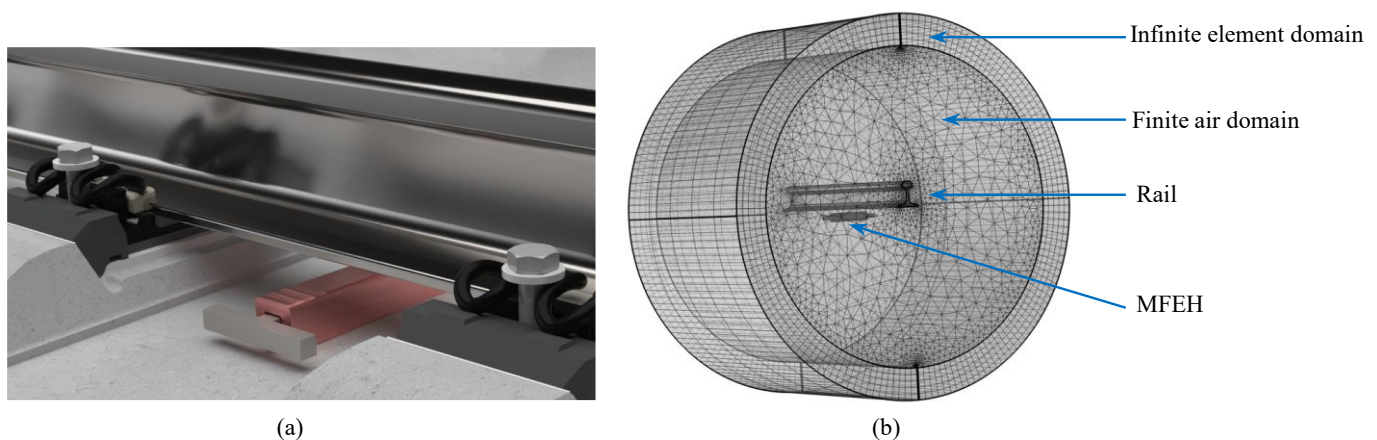


Figure 4. Magnetic field energy harvester (MFEH)-rail arrangement and the corresponding finite element model: (a) relative position of the MFEH and the rail; (b) finite element model of the MFEH.

Table 2. Material parameters of the COMSOL Multiphysics® finite element model [32].

Region	Material Description	Relative Permeability	Relative Permittivity	Electrical Conductivity (S/m)
Air domain	Air	1	1	1
Rail	54E1 rail steel	30	1	3×10^6
H-shaped core	Laminated silicon steel sheets (M7)	5000 [32]	1	4×10^4

The model consists of the rail, the MFEH, and the surrounding air domain. The rail adopts the standard 54E1 rail profile and has a length of 3 m. The air domain is represented by a cylindrical computational domain with a height of 3 m and a diameter of 2 m. The cylindrical geometry provides a finite exterior region around the rail–MFEH system and facilitates geometric coupling with the surrounding infinite element domain. To approximate the unbounded exterior space, an infinite element domain with a thickness of 0.2 m was added outside the finite air domain. Through a coordinate-stretching transformation, the infinite element domain reduces the artificial truncation error associated with applying a magnetic insulation condition at a finite distance from the region of interest [33]. The influences of the air-domain dimensions and the infinite-element thickness on the simulation results are further assessed in Section 3.3. The MFEH was positioned 50 mm below the rail bottom, as shown in Figure 4a, which was taken as the fixed installation distance in this parametric study so that the harvester remained within the relatively strong magnetic-field region beneath the rail. Material properties were assigned to each domain, as listed in Table 2.

The actual magnetic core of the MFEH is fabricated from laminated silicon steel sheets. To control the computational cost of the multi-parameter sweep, the laminated core was represented in the finite-element model as a homogenized magnetic body, while the suppression of eddy currents by interlaminar insulation was characterized through an equivalent electrical conductivity. This homogenization treatment and the electromagnetic-circuit-coupled modelling framework follow approaches previously validated against measurements of impedance, open-circuit voltage, and output power [19,22,32]. The present study builds on this framework to conduct the multi-parameter sweep analysis, and a representative validation case for the present finite-element model is provided in Section 3.3.

Both the rail and the H-shaped core are ferromagnetic materials, and their relative permeability may vary with the applied magnetic field. Previous MFEH studies have shown that a linear magnetization model with constant permeability is applicable when the magnetic-field strength and hysteresis loss are relatively low, as under the 50 Hz operating conditions considered here [19,25]. Therefore, both materials were treated using a linear magnetization approximation to reduce modelling complexity and improve computational efficiency. The constant relative permeabilities used for the rail and the laminated M7 core were adopted from the previous study [32]. This approximation is intended to evaluate the comparative effects of the investigated geometric and coil parameters rather than to reproduce the complete nonlinear magnetization, saturation, or hysteresis behavior of M7 steel.

Baseline geometry and parameter-sweep strategy: The baseline core geometry was taken as a center leg of $260 \times 20 \times 20 \text{ mm}^3$ ($l_c \times w_c \times h_c$) and side legs of $140 \times 20 \times 20 \text{ mm}^3$ ($l_s \times w_s \times h_s$). The coil was represented by an equivalent solid geometry to model the winding space and coupling region. The baseline coil span was 240 mm, with an outer cross-section of $50 \times 50 \text{ mm}^2$ and a central cavity of $25 \times 25 \text{ mm}^2$. A 2.5 mm manufacturing gap between the coil and the center leg was explicitly included in the finite-element ge-

ometry, and its influence is reflected in the numerical results. The radial winding thickness was 12.5 mm, and the number of turns was 6720. The core-geometry analyses followed a one-factor-at-a-time strategy, with all parameters other than the investigated core parameter maintained at their baseline values. Separate reference conditions were adopted for the coil-parameter analyses, as specified in Sections 4.3.1 and 4.3.2. In the analyses of coil span and number of coil turns, the target parameter variations were realized by adjusting the radial winding thickness in the equivalent representation. Therefore, the radial winding thickness in these two sets of cases did not remain at the baseline value.

The electromagnetic field was solved using the Magnetic and Electric Fields (MEF) interface with the magnetic vector potential–scalar electric potential (A–V) formulation, and the Coulomb gauge $\nabla \cdot \mathbf{A} = 0$ was imposed on the magnetic vector potential [34]. The outer boundary of the infinite element domain was assigned a magnetic insulation boundary condition. Considering that the model is symmetric with respect to the central plane, a perfect magnetic conductor boundary condition was applied on the symmetry plane, and only half of the model was solved to reduce the computational cost. A user-controlled meshing strategy was adopted. The infinite element domain was discretized using swept meshes, while the rail, the MFEH, and the finite air domain were meshed using tetrahedral elements. Local mesh refinement was applied in the vicinity of the rail and the core–coil region, whereas the remaining air domain was meshed more coarsely so as to balance accuracy and computational efficiency.

3.2. Simulation and Evaluation of MFEH Performance Metrics

The performance of the MFEH can be characterized in terms of both electrical output and equivalent-circuit parameters. The open-circuit voltage and the impedance-matched output power directly reflect the energy-harvesting capability of the device. The internal resistance and internal inductance determine the achievable power delivered to an external load. The open-circuit flux density characterizes the degree of magnetic-flux concentration inside the core under external excitation. The effective permeability reflects the equivalent response of the core structure and material to the external magnetic field. Accordingly, the present study adopts the open-circuit flux density, effective permeability, internal resistance, internal inductance, open-circuit voltage, and impedance-matched output power as key performance indicators to analyze the variation of MFEH performance under different conditions.

To evaluate these indicators, corresponding simulation cases were established in COMSOL Multiphysics®. First, the internal resistance and internal inductance of the MFEH were extracted from a frequency-domain study using the MEF interface, as shown in Figure 5a. In this simulation, the rail was de-energized, and a 0.01 A, 50 Hz current excitation was applied to the coil. The internal resistance and internal inductance were then obtained from the terminal voltage-current relationship of the coil. The extracted internal resistance therefore includes the winding resistance and the equivalent eddy-current resistance associated with eddy-current losses induced by the coil–current excitation in the conductive domains.

Second, the open-circuit voltage and the open-circuit flux density were evaluated with the coil open-circuited. As shown in Figure 5b, a sinusoidal AC current with an amplitude of 300 A and a frequency of 50 Hz was applied to the rail to simulate the traction-returning current, while the coil terminal was kept open. In the present study, the average magnetic flux density over the middle cross-section of the center leg was taken as the open-circuit flux density.

Finally, the output-power simulation incorporated the Electric Circuit interface to establish the external load circuit, as shown in Figure 5c. While the alternating current excitation was applied to the rail, the coil terminal of the MFEH, the tuning capacitor, and the

external load resistance were connected in series to form a closed circuit. Under the impedance-matching condition, the maximum output power corresponding to each operating case was obtained from the voltage and current across the load resistance. For each case, the tuning capacitance and load resistance were updated using the corresponding FEM-derived internal inductance and resistance to maintain resonance and impedance matching. Unless otherwise stated, the output power reported herein refers to the maximum output power under tuning and impedance-matching conditions.

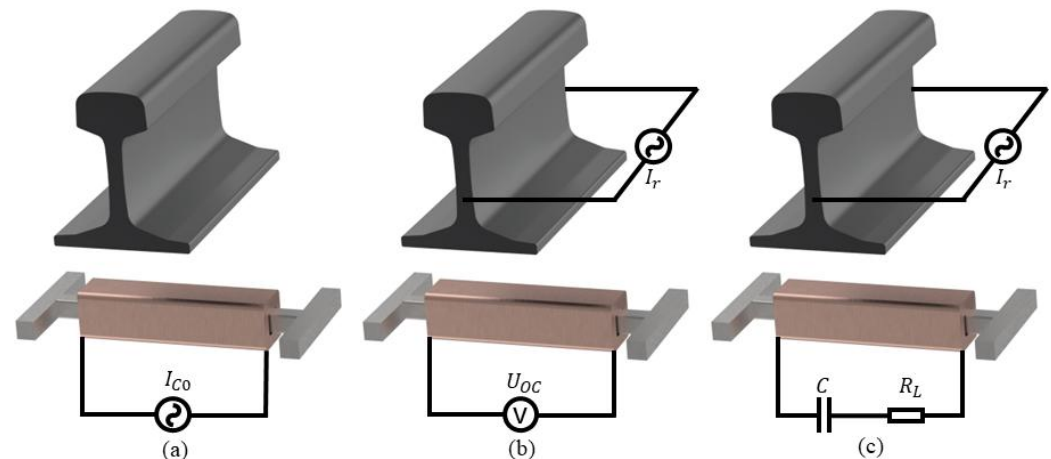


Figure 5. Finite element simulation configurations in COMSOL Multiphysics®: (a) extraction of internal inductance and internal resistance; (b) evaluation of open-circuit voltage and open-circuit flux density; (c) evaluation of impedance-matched output power.

3.3. Model Validation and Numerical-Domain Sensitivity Analysis

To assess model reliability, a representative validation case based on the prototype configuration reported in [22] was selected. Experimental measurements obtained in this study were compared with the simulation results. The validation device had a center-leg size of $260 \times 20 \times 20 \text{ mm}^3$ ($l_c \times w_c \times h_c$), a size of $140 \times 20 \times 20 \text{ mm}^3$ ($l_s \times w_s \times h_s$) for each side leg, a coil span of 240 mm, and 6720 coil turns.

As shown in Table 3, the relative errors in internal inductance, internal resistance, open-circuit voltage, and maximum output power were 3.70%, 4.83%, 3.83%, and 3.06%, respectively. All four errors were below 5%, with the largest observed for internal resistance. The agreement between the simulated and experimental results supports the predictive reliability of the developed finite-element model.

Table 3. Comparison between the experimental measurements and finite-element simulation results for model validation.

Performance Indicator	Experimental Value	Simulated Value	Relative Error
Internal inductance L (H)	13.77	13.26	3.70%
Internal resistance R (Ω)	202.60	192.81	4.83%
Open-circuit voltage U_{oc} (V)	51.98	53.97	3.83%
Maximum output power P_{max} (W)	1.96	1.90	3.06%

A numerical sensitivity analysis of the computational-domain parameters was subsequently conducted to evaluate the effects of the air-domain dimensions and infinite-element thickness on the calculated results. To provide a conservative assessment, a relatively large MFEH configuration within the investigated parameter range was selected, placing the device closer to the computational-domain boundaries than the baseline con-

figuration. The configuration had a center-leg size of $1500 \times 20 \times 20 \text{ mm}^3$ ($l_c \times w_c \times h_c$), a size of $1000 \times 20 \times 20 \text{ mm}^3$ ($l_s \times w_s \times h_s$) for each side leg, a coil span of 240 mm, and a radial winding thickness of 12.5 mm. The remaining device parameters were maintained at their baseline values.

For the baseline computational domain, the air-domain radius and height were 1000 mm and 3000 mm, respectively, and the infinite-element-domain thickness was 200 mm. Three comparison cases were generated using a one-factor-at-a-time approach by independently increasing the air-domain radius to 1500 mm, the air-domain height to 4500 mm, and the infinite-element-domain thickness to 300 mm. Only one computational-domain parameter was varied in each case, while all other geometric parameters, material properties, excitation conditions, and external-circuit settings were kept unchanged.

As shown in Table 4, increasing the air-domain radius, air-domain height, and infinite-element-domain thickness resulted in maximum relative differences among the four output quantities of less than 0.09%, 1.05%, and 0.05%, respectively. These small variations indicate that, within the investigated parameter range and the accuracy required in this study, further enlargement of the air domain or infinite-element domain has only a minor influence on the calculated results. The baseline computational-domain settings are therefore adequate for the present analysis.

Table 4. Numerical sensitivity analysis of the computational-domain settings of the finite-element model.

Case	Air-Domain Height (mm)	Air-Domain Radius (mm)	Infinite-Element-Domain thickness (mm)	L (H)	Relative Difference	R (Ω)	Relative Difference	U_{oc} (V)	Relative Difference	P_{max} (W)	Relative Difference
Baseline	3000	1000	200	34.863	—	1080.972	—	203.422	—	4.674	—
1	3000	1500	200	34.883	0.057%	1081.263	0.027%	203.492	0.034%	4.678	0.086%
2	4500	1000	200	35.229	1.050%	1090.994	0.927%	204.023	0.295%	4.710	0.770%
3	3000	1000	300	34.867	0.011%	1081.447	0.044%	203.428	0.003%	4.675	0.021%

Taken together, the experimental validation and computational-domain sensitivity analysis support the predictive reliability and numerical stability of the finite-element model within the investigated parameter range.

4. Parametric Effects on MFEH Performance

To quantitatively evaluate the power–volume trade-off associated with variations in the core geometry, the core–coil assembly volume V_{cc} and volumetric output power density P_v are introduced as auxiliary indicators. V_{cc} comprises the magnetic core and coil region, including the winding and the clearance required for winding and assembly, while P_v is defined as the impedance-matched output power per unit V_{cc} . The external circuit, mounting structure, protective enclosure, and other auxiliary components are excluded. These indicators characterize the power–volume performance of the MFEH rather than its complete structural cost.

4.1. Effect of Center-Leg Length

As the center leg is elongated, the demagnetizing effect is progressively weakened. Over the investigated center-leg length range of 260–1500 mm and under the same external flux density ($B_{ex} = 490.36 \mu\text{T}$), the open-circuit flux density and effective permeability both increase by approximately 180%, reaching 200.71 mT and 409.30, respectively, as shown in Figure 6a. Over the same range, the equivalent demagnetization factor decreases by 66.3% to 2.24×10^{-3} . The elongation increases the effective distance between the

equivalent magnetic poles at both ends of the core, thereby reducing the demagnetizing field and enhancing the magnetic-flux concentration capability [19,25].

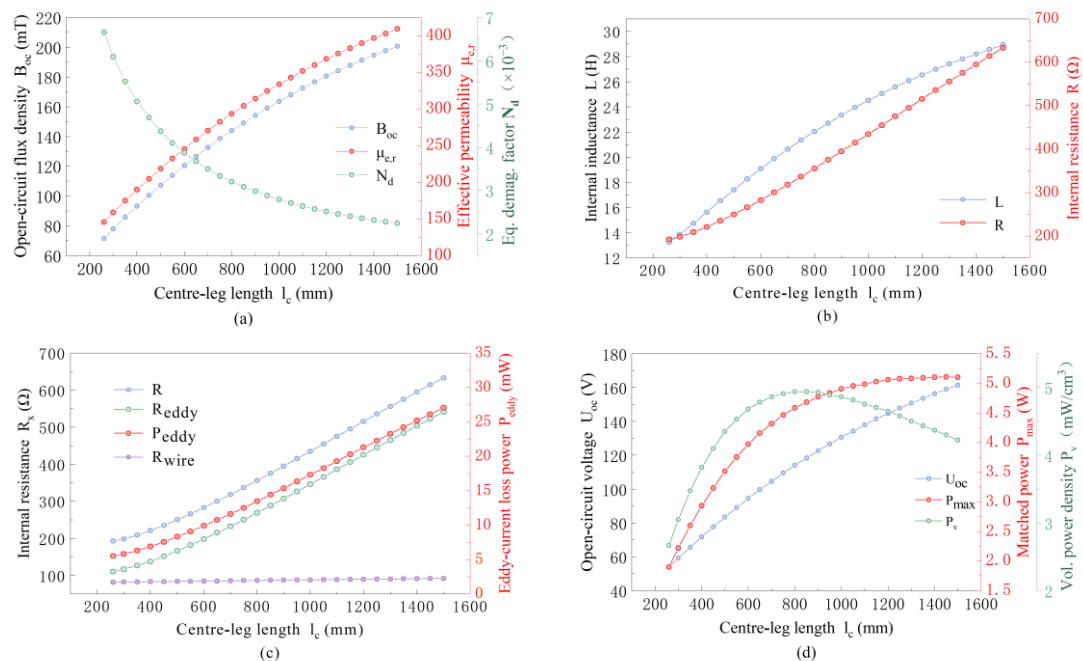


Figure 6. Effects of center-leg length l_c on the performance of the magnetic field energy harvester: (a) open-circuit flux density, effective permeability, and equivalent demagnetization factor; (b) internal inductance and internal resistance; (c) components of internal resistance and eddy-current loss power; (d) open-circuit voltage, impedance-matched output power, and volumetric output power density.

According to Equation (4), the increase in effective permeability leads to a corresponding increase in the internal inductance. As shown in Figure 6b, with the coil span and number of turns kept constant, the internal inductance increases monotonically as the center-leg length increases from 260 to 1500 mm, with an overall increase of 118.1%. Over the same parameter range, the internal resistance increases by 228.4%. The resistance decomposition shown in Figure 6c indicates that the equivalent eddy-current resistance increases to approximately 4.89 times its initial value, whereas the winding resistance increases by only 12.2%. Here, R_{eddy} is derived from the eddy-current loss directly calculated by the FEM and serves as its equivalent-circuit representation. Correspondingly, the total eddy-current loss power in the magnetic core and adjacent rail increases by approximately 388.7%. These comparisons indicate that the increase in internal resistance is primarily caused by the enhanced eddy-current effect associated with the increased magnetic flux density, whereas the contribution of the winding resistance remains relatively limited.

As shown in Figure 6d, the open-circuit voltage increases monotonically as the center-leg length increases from 260 to 1500 mm, reaching approximately three times its initial value. In contrast, the incremental gain in output power progressively diminishes. The output power reaches 4.91 W at a center-leg length of 1000 mm, representing an increase of approximately 158.5% relative to that at 260 mm. Further increasing the center-leg length from 1000 to 1500 mm provides only an additional power gain of approximately 4.0%. This diminishing gain occurs because the increase in internal resistance progressively offsets the benefit of the higher open-circuit voltage, causing the output power to approach a plateau of approximately 5.1 W.

The core–coil assembly volume V_{cc} increases linearly with the center-leg length. As the gain in output power gradually diminishes while the assembly volume continues to increase, the volumetric output power density P_v first increases and then decreases, reaching a maximum of 4.94 mW/cm^3 at a center-leg length of approximately 800 mm. As the center-leg length increases further from 800 to 1500 mm, the output power increases by only 11.3%, whereas V_{cc} increases by 30.2%, resulting in a 14.5% reduction in P_v . Accordingly, 800 mm is identified as the preferred numerical design point within the investigated range on the basis of power–volume performance. The larger center-leg lengths included in the parametric sweep are intended to reveal the saturation behavior rather than represent directly deployable dimensions. Although the 800 mm design imposes a smaller dimensional demand than the longer alternatives, its practical compatibility should still be evaluated according to the available installation space, installation orientation, sleeper spacing, ballast conditions, clearance limits, maintenance access, vibration-resistant mounting, and mechanical-protection requirements of the specific railway site.

4.2. Effects of Side-Leg Dimensions

4.2.1. Effect of Side-Leg Length

As the side-leg length increases from 20 to 1000 mm, the open-circuit flux density and effective permeability both increase by approximately 410.0%, while the equivalent demagnetization factor decreases by 81.8%, as shown in Figure 7a. The elongation of the side legs enhances the guidance of the surrounding external magnetic field, allowing more magnetic flux to be concentrated into the center-leg region. Meanwhile, it improves the magnetic-flux distribution at the core ends, thereby weakening the demagnetizing effect and reducing end leakage flux [19,28]. However, the changes in these quantities become progressively smaller at larger side-leg lengths, indicating diminishing improvement in magnetic-flux concentration.

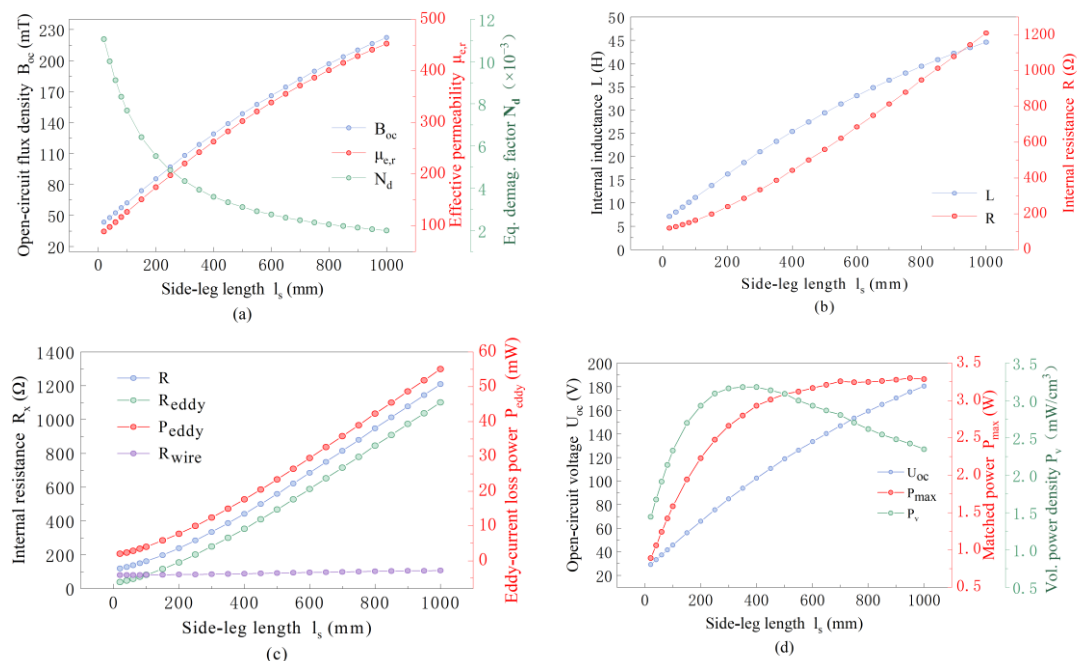


Figure 7. Effects of side-leg length l_s on the performance of the magnetic field energy harvester: (a) open-circuit flux density, effective permeability, and equivalent demagnetization factor; (b) internal inductance and internal resistance; (c) components of internal resistance and eddy-current loss power; (d) open-circuit voltage, impedance-matched output power, and volumetric output power density.

The internal inductance of the MFEH, shown in Figure 7b, and the open-circuit voltage, shown in Figure 7d, follow the same trend as the open-circuit flux density, both increasing with side-leg length. Over the investigated side-leg length range, the internal inductance increases from 7.13 to 44.61 H, while the internal resistance rises from 120.40 to 1210.27 Ω . The resistance decomposition shown in Figure 7c indicates that the equivalent eddy-current resistance increases by 1061.82 Ω , accounting for approximately 97.4% of the total increase in internal resistance. Correspondingly, the total eddy-current loss power in the magnetic core and adjacent rail increases from 2.01 to 55.10 mW. These results demonstrate that the increase in internal resistance is predominantly caused by the enhanced eddy-current effect.

As the side-leg length increases from 20 to 500 mm, the output power increases to approximately 3.06 W. Beyond 500 mm, the increase in internal resistance progressively offsets the benefit of the higher open-circuit voltage. As the side-leg length increases from 500 to 1000 mm, the open-circuit voltage increases by 51.9%, whereas the internal resistance increases by 116.1%. Substituting these FEM-derived values into Equation (10) predicts an output-power increase of approximately 6.8%, which is in close agreement with the 6.3% increase obtained from the electromagnetic-circuit-coupled simulation. This quantitative agreement indicates that the rapid increase in internal resistance, dominated by R_{eddy} , causes the output-power gain to progressively diminish. The output power approaches saturation at a side-leg length of approximately 700 mm and subsequently remains around 3.25 W.

From a power–volume perspective, the volumetric output power density P_v first increases and then decreases with increasing side-leg length, reaching a maximum of 3.18 mW/cm³ at approximately 400 mm, as shown in Figure 7d. As the side-leg length increases further from 400 to 1000 mm, the output power increases by only 12.2%, whereas the core–coil assembly volume V_c increases by 52.1%, resulting in a 26.3% reduction in P_v . Therefore, 400 mm is identified as the preferred numerical design point within the investigated range, as it maximizes the power–volume performance while limiting the dimensional increase associated with longer side legs. The upper bound of 1000 mm for the side-leg length was selected to characterize the transition from rapid power growth to saturation rather than to represent a directly deployable engineering dimension.

4.2.2. Effects of Side-Leg Height and Width

As the side-leg height increases, the effective cross-sectional area available for magnetic-flux concentration increases, while the end demagnetizing effect is weakened with increasing cross-sectional area. Consequently, the open-circuit flux density and effective permeability increase approximately linearly, as shown in Figure 8a [28,35]. When the side-leg height increases from 20 to 200 mm, both quantities increase by approximately 88.0–121.8% over the investigated side-leg widths. Meanwhile, increasing the side-leg width also enhances the open-circuit flux density and effective permeability. For example, at a side-leg height of 100 mm, increasing the side-leg width from 20 to 100 mm increases both quantities by approximately 49.1%. This is because increasing the side-leg width along the axial direction of the center leg extends the axial range over which the end flux-collecting structure acts, thereby improving the magnetic-flux distribution at the ends and weakening the demagnetizing effect [28]. The mechanism associated with increasing the side-leg width is therefore similar to that of increasing the side-leg length.

As shown in Figure 8b–d, the open-circuit voltage, internal resistance, and output power all increase with side-leg height. As the side-leg height increases from 20 to 200 mm, the open-circuit voltage increases by 94.5–138.4% and the internal resistance increases by 190.8–223.6% over the investigated side-leg widths, while the corresponding output-power increase ranges from 28.0% to 74.2%. Similarly, increasing the side-leg

width raises the open-circuit voltage, internal resistance, and output power of the MFEH, exhibiting trends consistent with those observed for the side-leg height.

From an engineering perspective, increasing the side-leg height enlarges the vertical envelope, whereas increasing the side-leg width increases the longitudinal dimension and material usage. As h_s increases from 20 to 200 mm, P_v decreases by 27.9–76.0% over the investigated widths, indicating that the increase in assembly volume exceeds the output-power gain. The effect of w_s depends on h_s : increasing w_s from 20 to 100 mm raises P_v by 13.8% at $h_s = 20$ mm but reduces it by 49.1% at $h_s = 100$ mm. Among the investigated combinations, the maximum P_v of approximately 3.00 mW/cm³ is obtained at $h_s = 20$ mm and $w_s = 100$ mm. Therefore, a relatively small side-leg height is preferable, and the side-leg width may be moderately increased when the height is maintained at a low value.

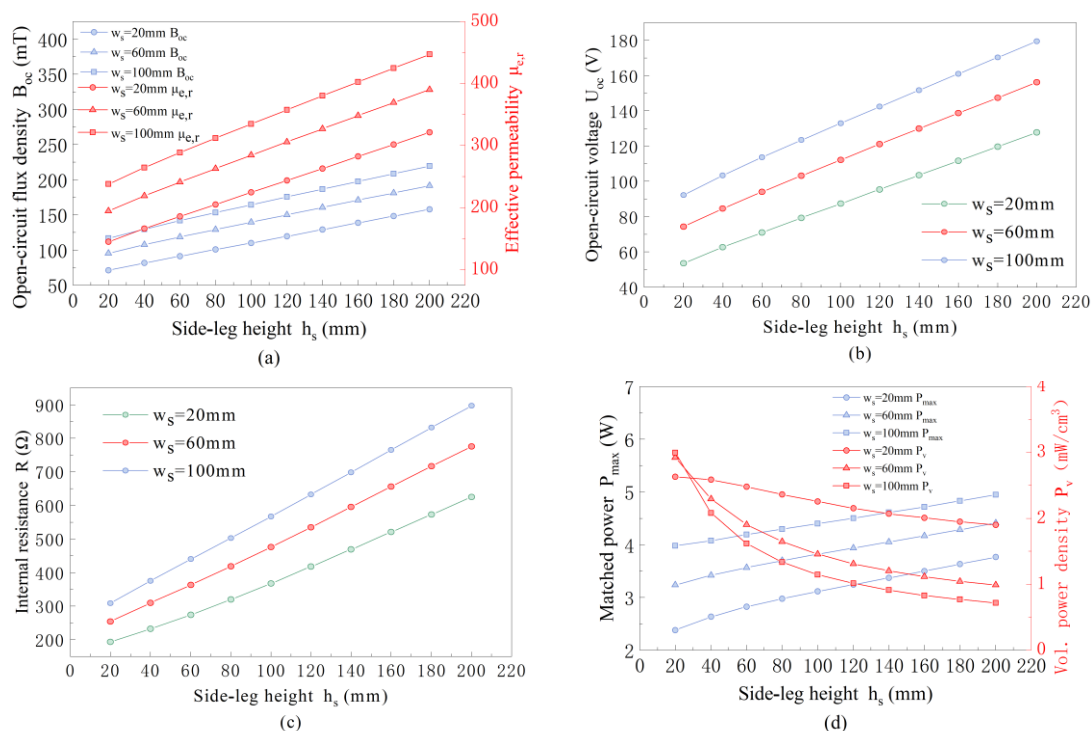


Figure 8. Effects of side-leg height h_s and width w_s on the performance of the magnetic field energy harvester: (a) open-circuit flux density and effective permeability; (b) open-circuit voltage; (c) internal resistance; (d) impedance-matched output power and volumetric output power density.

4.3. Effects of Coil Parameters

4.3.1. Effect of Coil Span

To analyze the effect of the coil span l_{coil} on the performance of the MFEH, the number of coil turns was fixed at 3360. The number of winding layers, and hence the radial winding thickness, was adjusted as l_{coil} varied from 40 to 250 mm.

As shown in Figure 9a, the open-circuit flux density at the middle cross-section of the center leg and the effective permeability vary by less than 0.4% over the investigated coil-span range and therefore remain essentially unchanged. This indicates that these quantities are determined mainly by the core geometry and external magnetic field rather than by the coil span. Varying the coil span has a different physical meaning from varying the center-leg length. Changing the center-leg length directly alters the core geometry, demagnetizing effect, and magnetic-flux concentration capability. By contrast, changing the coil span only modifies the coverage of the winding on the existing center leg without

altering the core geometry. Under ideal open-circuit conditions, the coil current is zero and therefore does not perturb the magnetic-field distribution.

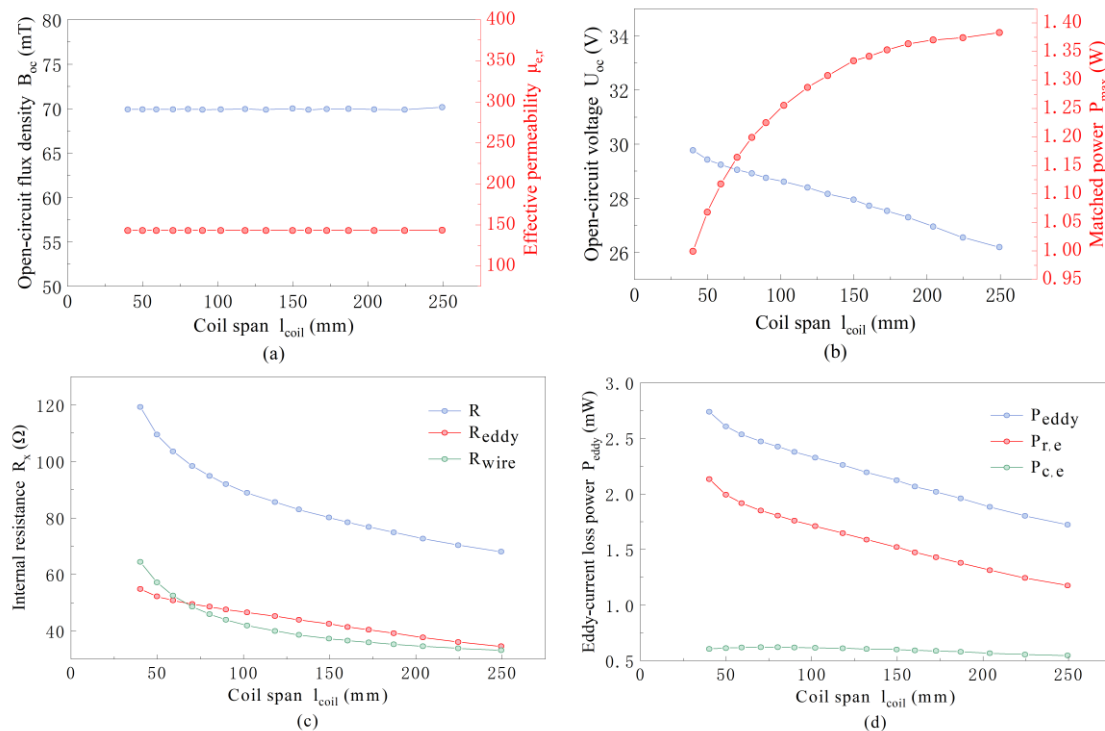


Figure 9. Effects of coil span on the performance of the magnetic field energy harvester: (a) open-circuit flux density and effective permeability; (b) open-circuit voltage and impedance-matched output power; (c) internal resistance and its components; (d) eddy-current loss power and its components.

Despite the nearly unchanged open-circuit flux density, the open-circuit voltage decreases monotonically from 29.77 to 26.20 V as the coil span increases from approximately 40 to 250 mm, corresponding to a reduction of 12.0%, as shown in Figure 9b. This is because the magnetic flux density in the center leg is distributed non-uniformly along the axial direction, reaching its maximum near the middle and decreasing toward both ends. As the coil span increases, the winding extends from the central high-flux-density region toward the lower-flux-density regions at both ends, reducing the average magnetic flux linked by each turn and thereby lowering the induced voltage [25,28].

Although the open-circuit voltage decreases, the output power increases from approximately 1.00 to 1.38 W because the internal resistance decreases by 43.0%, from 119.34 to 68.08 Ω, as shown in Figure 9b,c. With the total number of turns kept constant, increasing the coil span reduces the radial winding thickness, thereby decreasing the average length per turn, total wire length, and winding resistance. Meanwhile, under the same coil-current excitation, increasing the coil span distributes the coil-generated magnetic field over a longer axial range, thereby reducing the local magnetic flux density in the core and the eddy-current density induced in the rail, as shown in Figure 10. Consequently, the winding resistance and equivalent eddy-current resistance decrease by 48.6% and 37.1%, respectively, as shown in Figure 9c. The reduction in R_{eddy} is mainly associated with the 44.8% decrease in rail eddy-current loss, whereas the core eddy-current loss decreases by only 9.7%, as shown in Figure 9d. Therefore, the reduction in internal resistance outweighs the adverse effect of the decreasing open-circuit voltage, resulting in an overall increase in output power. When l_{coil} increases further from approximately 200 to 250 mm, the output power increases by only 0.9%, indicating that it gradually approaches a plateau, as shown in Figure 9b.

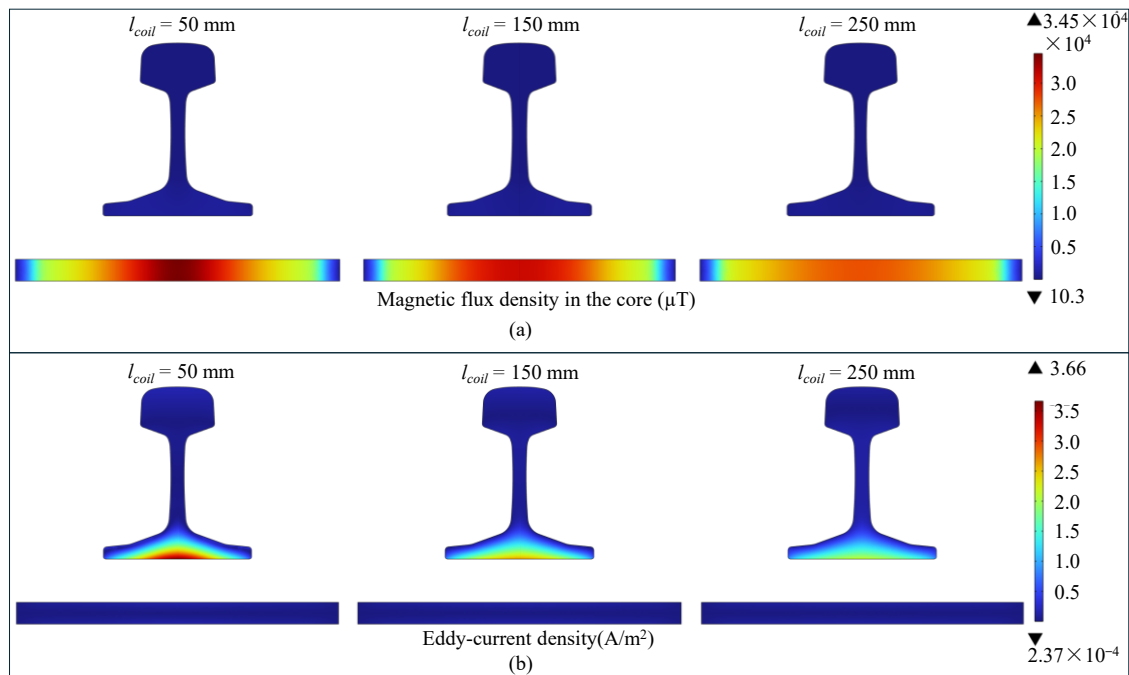


Figure 10. Effects of coil span under a 0.01 A, 50 Hz coil-current excitation with the rail de-energized: (a) magnetic flux density distribution in the core; (b) eddy-current density distribution in the rail.

4.3.2. Effect of the Number of Coil Turns

To analyze the effect of the number of coil turns N on the performance of the MFEH, the coil span l_{coil} was fixed at 240 mm, while the radial winding thickness was varied so that N ranged from 1440 to 20,160. As noted in Section 4.3.1, the open-circuit flux density varies negligibly with the coil parameters and therefore remains essentially unchanged.

According to Equation (1), the open-circuit voltage increases approximately linearly from 11.45 to 163.88 V as the number of coil turns increases from 1440 to 20,160, as shown in Figure 11a. Meanwhile, the output power increases from 0.91 to 2.46 W, although the power gain progressively diminishes. When the number of turns doubles from 10,080 to 20,160, the open-circuit voltage increases by 102.7% and the internal resistance by 253.1%, whereas the output power increases by only 16.5%. This behavior is mainly caused by the rapid increase in internal resistance.

As shown in Figure 11b, the increase in internal resistance R , from 18.05 to 1382.54 Ω , originates from two sources. First, increasing the number of turns increases the total wire length, causing the winding resistance R_{wire} to increase from 12.87 to 402.12 Ω . Second, under the same coil-current excitation, increasing the number of turns strengthens the electromagnetic excitation imposed on the core and adjacent rail. Consequently, the rail and core eddy-current losses increase from 0.15 and 0.11 mW to 29.49 and 19.53 mW, respectively, raising the total eddy-current loss power P_{eddy} from 0.26 to 49.02 mW. Accordingly, the equivalent eddy-current resistance R_{eddy} increases from 5.18 to 980.42 Ω and accounts for approximately 70.9% of the total internal resistance at $N = 20,160$. These results indicate that the rapid increase in internal resistance, dominated by the equivalent eddy-current resistance, progressively limits the output-power gain obtained by increasing the number of coil turns.

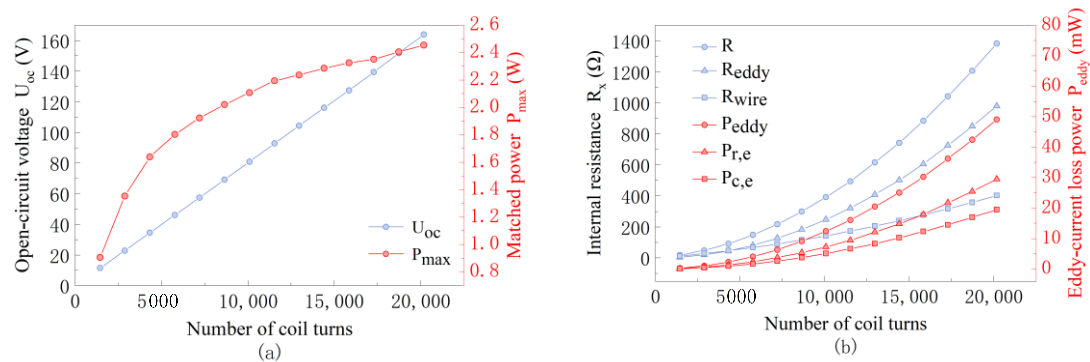


Figure 11. Effects of the number of coil turns on the performance of the magnetic field energy harvester: (a) open-circuit voltage and impedance-matched output power; (b) components of internal resistance and eddy-current loss power.

5. Conclusions

To address the lack of systematic parametric design rules for H-shaped-core MFEHs under railway traction-returning magnetic fields, a three-dimensional electromagnetic-circuit-coupled finite-element model including the rail, the MFEH, and the air domain was established. The model was validated against experimental measurements, with relative errors in the principal performance indicators below 5%. The effects of core geometry and coil parameters on the magnetic-circuit characteristics and electrical output performance were systematically investigated. The main conclusions are as follows:

- (1) Increasing the lengths of the center leg and side legs weakens the demagnetization effect and enhances the open-circuit flux density and effective permeability but also intensifies eddy-current losses in the core and the adjacent rail, resulting in a rapid increase in internal resistance. Consequently, the incremental gain in output power progressively diminishes, and the output power approaches saturation. Under the operating conditions considered in this study, when the center-leg length exceeds approximately 1000 mm, the output power approaches 5.1 W; when the side-leg length exceeds approximately 700 mm, the output power approaches 3.25 W. From the perspective of power–volume performance, the volumetric output power density reaches its maximum at center-leg and side-leg lengths of approximately 800 and 400 mm, respectively. These dimensions are therefore identified as the preferred numerical design points within the investigated ranges.
- (2) Increasing either the side-leg height or the side-leg width enhances the magnetic-flux concentration capability and improves the electrical output of the MFEH. Within the investigated range, the open-circuit flux density and effective permeability increase approximately linearly with side-leg height. However, enlarging the side-leg dimensions also increases the core–coil assembly volume, and the resulting output-power gain does not necessarily compensate for the accompanying increase in volume. Therefore, the side-leg height and width should be selected by balancing electrical output, volumetric output power density, structural compactness, and railway installation constraints. A relatively small side-leg height is preferable, while the side-leg width may be moderately increased when the height is maintained at a low value.
- (3) The coil parameters require a trade-off between voltage gain and loss. With the number of turns kept constant, increasing the coil span reduces the open-circuit voltage but also lowers the winding resistance and equivalent eddy-current resistance so that the output power still increases and gradually approaches a plateau. Increasing the number of coil turns raises the induced voltage approximately linearly, but the accompanying increases in winding resistance and equivalent eddy-current resistance

progressively offset this gain, causing the output power to increase rapidly at first and then more slowly. Therefore, the coil span and the number of turns should be determined jointly according to the target output voltage, allowable loss level, and available winding space.

In summary, the design of the H-shaped-core MFEH is essentially a multiphysics-coupled optimization problem involving magnetic-flux concentration, demagnetization, eddy-current loss, winding resistance, assembly volume, and circuit response. The structure-loss-performance relationships revealed in this study provide a quantitative basis for the parametric design and engineering application of self-powered devices for low-power wireless monitoring nodes in railway environments.

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