

Article

Electromagnetic Design and Analysis of a Stator–Magnet Transverse Flux Linear Oscillatory Machine with Yokeless Mover Core

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Abstract: Conventional stator–magnet moving–iron transverse–flux linear oscillatory machines (CSMTLOMs) are widely applied in directly–drive reciprocating devices due to the merits of easy fabrication and robust mover. However, in order to keep the mover vibrating at a certain resonance frequency to save the energy and enlarge the output power, they still suffer from a higher requirement on spring stiffness due to their thick and heavy mover core, which would also narrow the frequency band with a high power factor due to the large inertial energy storage of the heavy mover. Hence, to reduce the mover core weight to reduce the demand of the spring and improve the operation performance, an improved linear oscillatory machine featured by a spoke–type interior permanent magnet inner stator (ISMTLOM) is proposed. Benefiting from its separated two stators, the tangential flux in the radial plane can return through the inner stator core, so that the yoke of the mover core can be eliminated directly. Then, to analytically investigate the influence of the special axial local saturation effect, the segmental equivalent magnetic circuit (EMC) model of the ISMTLOM is established, wherein a saturation coefficient is introduced to quantitatively consider the local saturation effect on the output force. Consequently, several important size parameters are optimally selected when keeping the same outer diameter and copper loss as that of the CSMTLOM. Afterward, the three–dimension finite element algorithm (3D FEA) is adopted for the electromagnetic performance validation and comparison. Finally, it is found that the nonlinear segmental EMC corrected by the saturation coefficient can quickly predict the output force more accurately within the wide load range, and benefiting from the topology improvement, the ISMTLOM has the merits over the CSMTLOM in its smoother output force, much lighter mover core, and less demand of mechanical spring stiffness, whilst preserving the similar output force density.

Keywords: linear oscillatory machine; mover core weight; saturation effect; equivalent magnetic circuit (EMC); three–dimension finite element algorithm (3D FEA)



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1. Introduction

Permanent magnet linear oscillatory machines (PMLOMs) are promising when they are used for directly driving linear reciprocating devices, such as combined heat and power systems, Stirling free piston generation, and vibration energy harvesters, for its direct–drive transmission mode, compact structure, and high force density [1–4]. In practical application, PMLOMs usually work with several pairs of mechanical springs to form a force harmonic vibration system [5], as shown in Figure 1, which not only can automatically reset the mover back to balance position, but also can utilize the mechanical resonance to amplify the stroke [6]; the higher the resonance frequency is, the larger the output power is. In this way, PMLOMs working in the mechanical resonance can achieve a large displacement with a rather small current, which is beneficial for saving energy [7]. Hence, as to PMLOMs, besides the electromagnetic performance, the mover weight should

also be considered due to its significant influence on the resonance and the spring stiffness requirement.

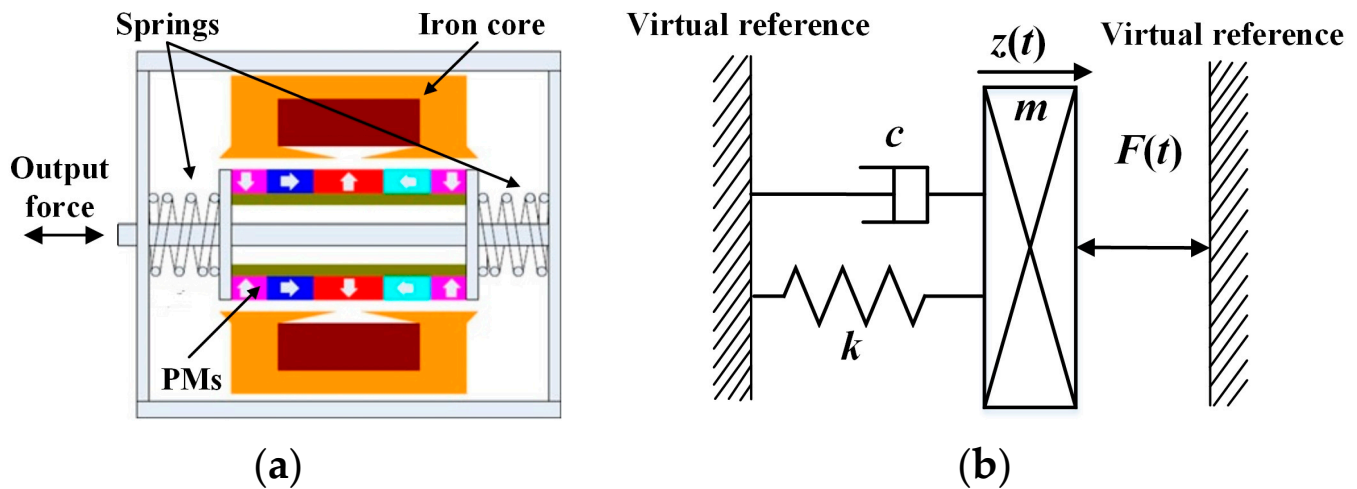


Figure 1. Completed PMLOM drive in practical application. (a) Overall sketch of the PMLOM drive. (b) Equivalent harmonic vibration system, where $F(t)$ is the dynamic force, $z(t)$ the mover transient displacement, k the spring stiffness, and c the damping coefficient.

In early research, the planar structure is the commonly used topology, which is easy for core lamination and assembly [8]. These planar PMLOMs can be classified by the surface-mounted permanent magnet and the interior permanent magnet mover [9]. The surface-mounted permanent magnet mover has the advantage of easy fabrication while having the disadvantages of a larger air gap and difficulty in protecting the magnets. The interior permanent magnet mover can concentrate the flux to enhance the force density, but it still suffers from high detent force due to the severe cogging effect [10]. The common defect of the planar PMLOM is that the overall volume is too large because of the long mover [11]. It makes them not applicable for some reciprocating devices in which space is rather limited.

To reduce the overall volume, some researchers have sought for compacter alternative topologies, in which the tubular structure is preferred [12]. Generally, the flux path of the traditional tubular PMLOM locates in the same plane as the moving direction [13]. In this case, to reduce the iron loss, its stator core must consist of either circumferentially laminated silicon steel [14], or a solid core made of soft magnetic composite [14–19]. The former choice would lead to a lower stacking factor, higher flux leakage [14], and higher fabrication difficulty and cost [20], while the latter one would cause lower force density and higher magnetization loss due to the lower permeability of soft magnetic composite material [21]. Therefore, there is an urgent need to switch the traditional way of lamination to construct a tubular core of LOMs, as well as the design method of the flux path. Moreover, in traditional topologies, the permanent magnets (PMs) are usually placed on the mover, which vibrate at high frequency and thus easily suffer from the mechanical damage [22]. To protect the PMs, these traditional topologies always require a high-strength non-ferromagnetic bracket, which increase the fabrication difficulty and reduce the reliability [23].

For continued adoption of laminated silicon steel as the core material and to simplify the stacking process, some researchers have introduced the transverse flux structure as the flux path design, and developed moving-magnets-type LOMs with double outer stators [24,25]. Although changing the flux path design can make the lamination method the same as the traditional rotary electrical machines, the PMs of these topologies are still fixed on the mover, resulting in high mechanical stress in the PMs during the reciprocating motion. Therefore, to further protect the PMs from vibration, some researchers firstly propose dual-stator transverse flux LOMs with yoke-inserted PMs (CSMTLOM) in [26], which is further fully analyzed in [27], as depicted in Figure 2a,b. It adopts the similar

axial-laminated core as the traditional rotary-switched reluctance machine, which can reduce the fabrication cost and enhance the mover strength. Full investigation has demonstrated that whilst having the merit of simpler lamination, it can reach a similar output force and efficiency as the prevailing moving-magnet-type LOM under the same size limitation [28]. Nevertheless, this structure still suffers from a heavy mover core, because a relative thick mover yoke needs to be retained as the returning path of the tangential flux [29]. In this case, a pair of moderately rugged mechanical springs is required to reach the same resonant frequency, which increases the overall cost and the assembly difficulty. Additionally, the moving mass usually works as an inductance in the mechanical system to store the kinetic energy during the vibration. Hence, if the driving frequency deviates from the resonance frequency during the operation, the large moving mass would easily decrease the power factor, which is the other reason why PMLOMs need to reduce the mover core weight.

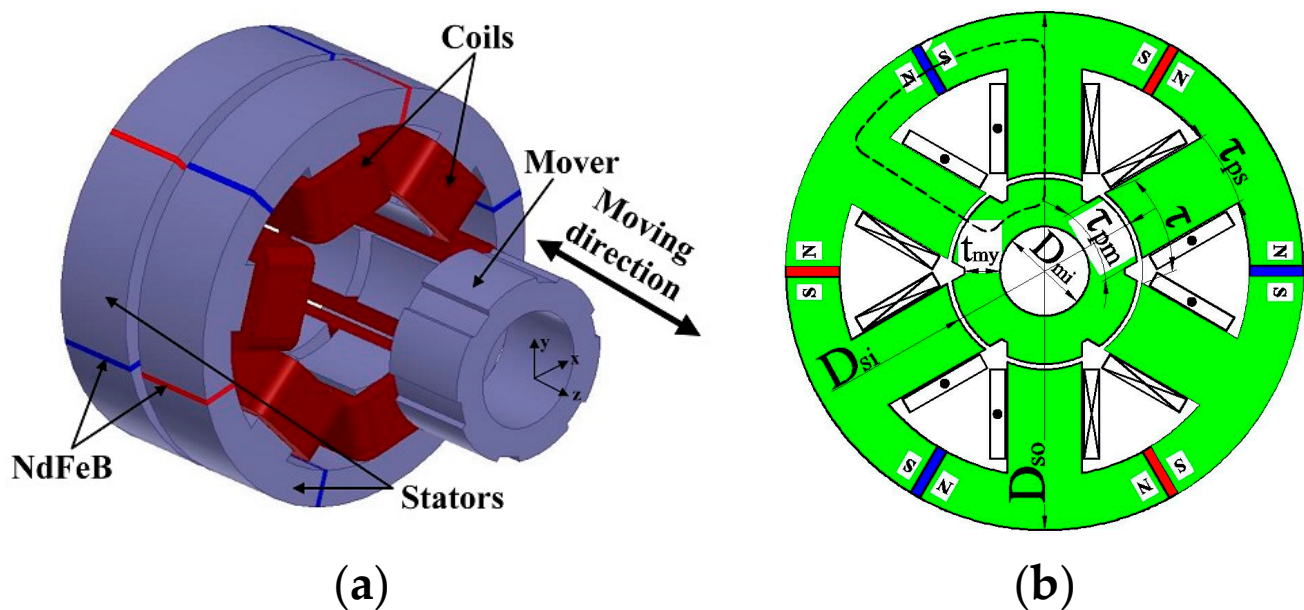


Figure 2. Topology of the CSMTLOM: (a) 3D scheme; (b) radial cross-sectional view.

Furthermore, based on the developed topologies, different calculation models for performance analysis have been established, which can be classified into three categories. The most commonly used one is the three-dimension finite element algorithm (3D FEA), which is most accurate but also time-consuming [30]. As an alternative, some researchers have resorted to the subdomain method to relieve the calculation burden, which can quickly calculate the no-load performance with satisfactory accuracy [31]. Other researchers have adopted the lumped parameters magnetic circuit to calculate the electromagnetic performance, which also can provide good agreement with the 3D FEA under light load conditions [32]. However, these two fast analytical models cannot manage the field calculation for the dual-stators transverse-flux PMLOMs, especially under the overload condition. It is because the total stator length of these types of machines is longer than the mover, leading to a nonuniform flux distribution along the moving direction when considering the armature reaction, which is the so-called special axial local saturation effect. In this case, the flux density in different axial areas should be distinguished and modeled separately according to the saturation extent. Otherwise, large errors would occur during the calculation.

In general, although many researchers have developed various stator–magnet moving–iron transverse–flux LOMs as the robust drive for reciprocating devices, the common issue of a heavy mover core needs an urgent solution. Moreover, still lacking is an effective analytical method to manage the axial local saturation effect to quickly and accurately calculate the output force within a wide load range. Therefore, one inner stator–magnet moving–iron transverse–flux LOM (ISMTLOM) with two separated stators is proposed, of which PMs are inserted in the inner stator yoke as the spoke type. It separates the windings and PMs on different side stators, which can not only still prevent the PMs from vibration, but also can significantly reduce the mover core thickness and its weight. In Section 2, this improved topology and its operation principle are illustrated in detail. In Section 3, analytical modeling of the ISMTLOM is carried out by the equivalent magnetic circuit (EMC), including the discussion concerning its special local saturation effect. Then, in Section 4, by selecting the most important size parameters, the optimal design of the ISMTLOM is obtained, and its main performance verified and compared with the CSMTLOM, including both the electromagnetic performance and the operation indices in the overall control system. Finally, the conclusions are provided in Section 5.

2. Topology Improvement and Operation Principle

2.1. New Topology Description

As can be seen in Figure 3a,b, the proposed ISMTLOM has an improved mover and PM arrangement when compared with the CSMTLOM. Its stators are separated into inner and outer parts, and the outer part has two cores adopting the similar structure as the switched reluctance machine, with armature coils wound around two axial–aligned teeth. Its inner stators are in cylinder shape, with spoke–type magnets inserted in it. Similar to the CSMTLOM, the arrays of PMs on the different inner stators are opposite, to gain the reversed PM magnetomotive force (MMF) on different sides. In this scheme, the PMs would stay away from the winding (main heat source), which is located on the outer stators, and thus gain a lower temperature rise at the rated condition. Furthermore, the mover is installed in the air gap between the inner and outer stators; as can be seen from Figure 4d, the air gap can be divided into Area I and II according to different lengths. The length of the former is longer, and the latter is much shorter. Because the mover consists of thin ring–type iron cores only, the mover weight can be greatly reduced.

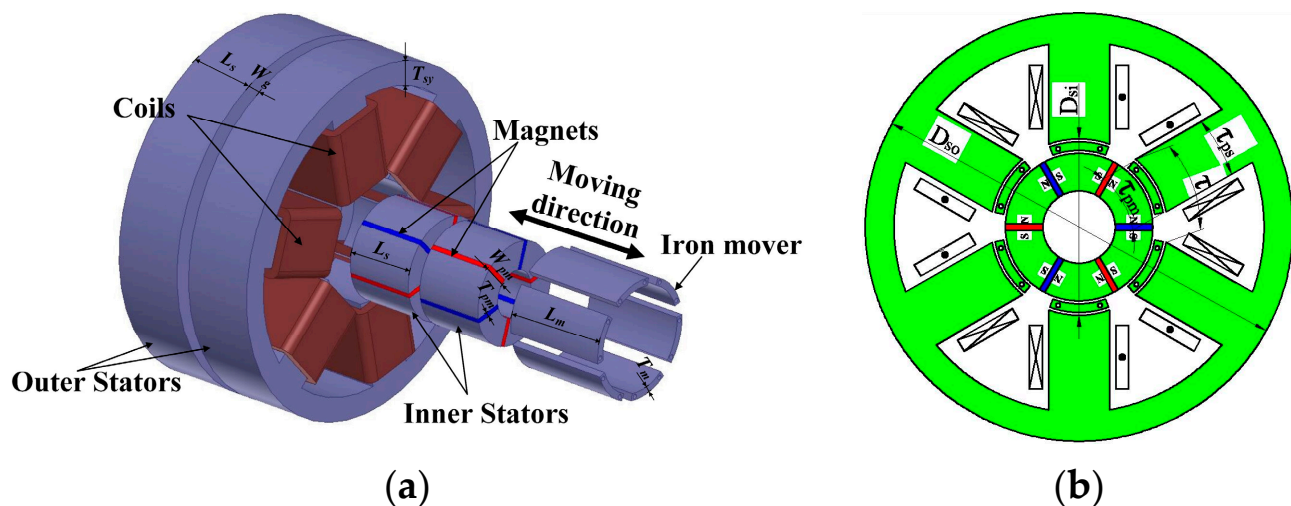


Figure 3. Topology of the ISMTLOM: (a) 3D scheme; (b) radial cross–sectional view.

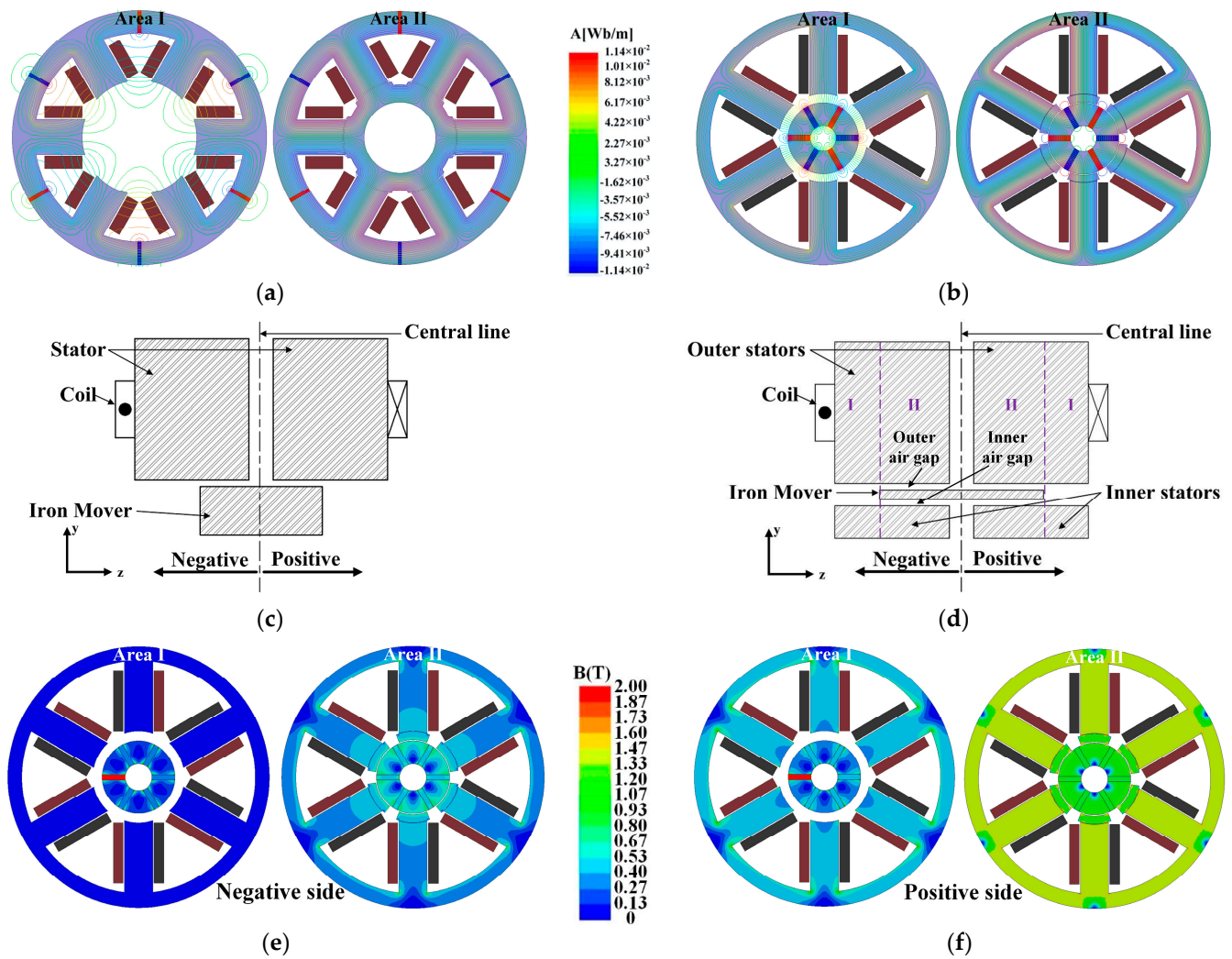


Figure 4. Flux path, axial cross-sectional view, and flux density distribution. (a) Flux path of the CSMTLOM. (b) Flux path of the ISMTLOM. (c) Axial cross-sectional view of the CSMTLOM. (d) Axial cross-sectional view of the ISMTLOM. (e) Flux density of the ISMTLOM in its negative side under the rated load condition. (f) Flux density of the ISMTLOM in its positive side under the rated load condition.

2.2. Operation Principle

As can be seen in Figure 4a–d, the same as that of the CSMTLOM, the ISMTLOM also moves along the axial direction. Hence, its moving trajectory is perpendicular to the flux in the radial plane. When the winding is excited by a dc current, the armature MMF drops on its two inner stators are the same. However, because of the opposite PM MMFs on the two inner stators, the magnetic field intensity in the two air gaps in the different sides are unbalanced, as shown in Figure 4e,f. As illustrated in Figure 4d, when the mover aligns with the central line, the armature MMF weakens the field in the negative side while strengthens the field in positive field, and consequently the iron mover would be driven positive towards, and vice versa.

In practical operation, a single-phase sinusoidal current is flow in the armature winding. So, within one driving cycle, the alternating armature MMF would periodically strengthen and weaken the field in the negative and positive side, and as a result, the intensity of the air-gap field in the different sides would also change periodically. Hence, the mover would be driven back and forth along the axial direction by the dynamic force. In this case, with the aid of mechanical springs set at the double side of the mover, the ISMTLOM would become a forced harmonic system to achieve a larger stroke with a

smaller dynamic force, which means lower winding current and less power loss, and therefore beneficial for saving energy.

3. Analytical Modeling of the ISMTLOM

To analyze the electromagnetic performance of the ISMTLOM analytically, the segmental equivalent magnetic circuit (EMC) model is adopted to reveal the quantitative relation between the main size parameters and the output performance. The magnetic fields in the ISMTLOM are mainly excited by both PMs and armature current, but some of their flux does not share the same paths. Therefore, it is better to build the EMC of them separately, and then further consider the interaction where their flux paths overlap. It should be noted that the flux density in Area I and Area II is at a different level, because the total air-gap length (δ_1) in Area I is much longer than that in Area II (δ_{2i} for inner air gap and δ_{2o} for outer air gap). As a result, the superposition law can be directly applied to Area I. But in Area II, the saturation effect should be considered when the PM and armature fields are mutually positively strengthened, which only occurs in one side in Area II within the half stroke. This phenomenon can be illustrated more specifically by Figure 4. As can be seen in this figure, the air-gap flux density is apparently much higher in Area II due to the shorter air-gap length. So, when the direction of the armature field in Area II is the same as that of the PM field, the iron core easily suffers from a saturation effect.

3.1. EMC of No-Load Condition (Excited by PMs Only)

According to Figure 5, when only considering the magnetic field excited by PMs, the EMC of these two areas can be depicted as shown in Figure 5, where Φ_r is the remanence flux, $R_{\delta_{2i}}$ and $R_{\delta_{2o}}$ are the inner and outer air-gap reluctance in Area II, $R_{\delta_{1l}}$ is the leakage reluctance between each of the two adjacent mover cores in Area II, R_{δ_1} is the air-gap reluctance in Area I, and A_{m1} and A_{m0} are the leakage permeance and self-demagnetization permeance of PMs. These parameters can be calculated by the expressions given in Appendix A.

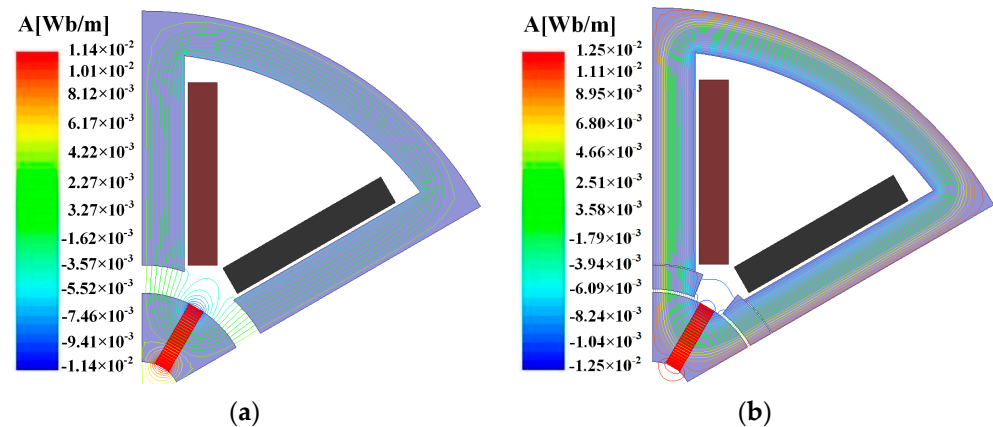


Figure 5. PM flux path: (a) Area I; (b) Area II.

Then, according to Figure 6a, the air-gap flux density in Area I can be calculated by

$$B_{\delta_1} = \frac{B_r W_m N_p}{\pi k_{so} [1 + 2R_{g1} (A_{m0} + A_{m1})] (D_{si} - \delta_1)} \quad (1)$$

where N_p is the number of poles and equal to six for the ISMTLOM. Similarly, the inner and outer air-gap flux density in Area I and II can also be derived as follows:

$$B_{\delta_{2i}} = \frac{2N_p B_r W_m}{[1 + (2R_{\delta_{2i}} + \frac{2R_{\delta_{2o}} R_{\delta_{1l}}}{2R_{\delta_{2o}} + R_{\delta_{1l}}}) (\Lambda_{m0} + \Lambda_{m1})] [\pi (D_{si} - 2\delta_{2o} - 2W_{iron}) k_{so} k_{ms} + 2N_p \delta_{2i}]} \quad (2)$$

$$B_{\delta 2o} = \frac{2N_p B_r W_m R_{il}}{[1 + (2R_{\delta 2i} + \frac{2R_{\delta 2o} R_{il}}{2R_{\delta 2o} + R_{il}})(A_{m0} + A_{ml})](2R_{\delta 2o} + R_{il})[\pi D_{si} k_{so} + 2N_p \delta_{2o}]} \quad (3)$$

where k_{so} is the slot open coefficient, which is defined by $k_{so} = \tau_{ps}/\tau$, and k_{ms} is the pole arc of mover iron, which is defined by $k_{ms} = \tau_{pm}/\tau_{ps}$.

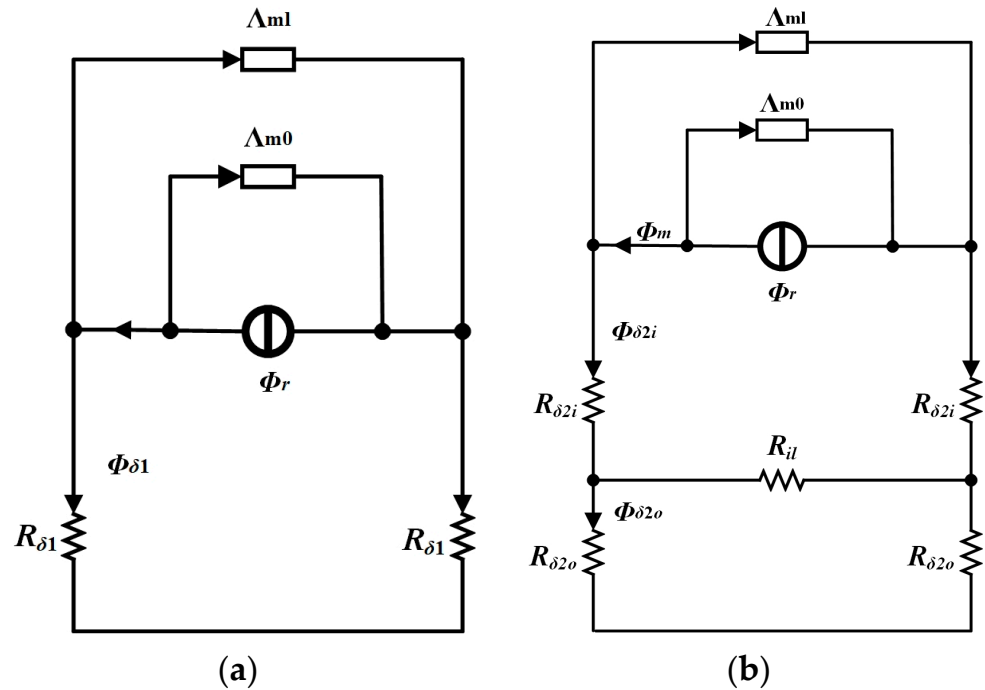


Figure 6. EMC under the no-load condition: (a) Area I; (b) Area II.

3.2. EMC of Armature Reaction (Excited by Winding Current Only)

When only considering the field generated by the winding current, the flux line distribution in Area I and II is shown in Figure 7. Referring to Figure 7, it can be seen from this figure that the armature flux would cross the slot area, where the PM flux would not appear. Hence, the slot leakage flux should be considered in the EMC of the armature reaction.

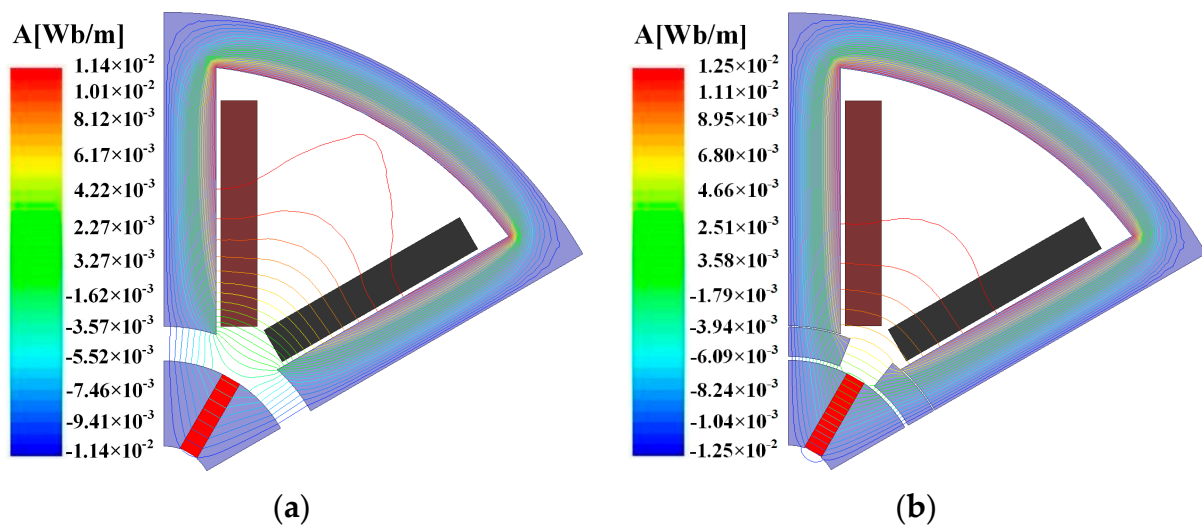


Figure 7. Armature flux path: (a) Area I; (b) Area II.

According to Figure 7, the EMC of the armature reaction can be depicted as shown in Figure 8, where F_a is the magnetomotive force of winding, R_{sl} is the slot leakage reluctance, and R_{ml} and R_{m0} are the reciprocals of Λ_{ml} and Λ_{m0} . The expressions of parameters are also listed in Appendix A.

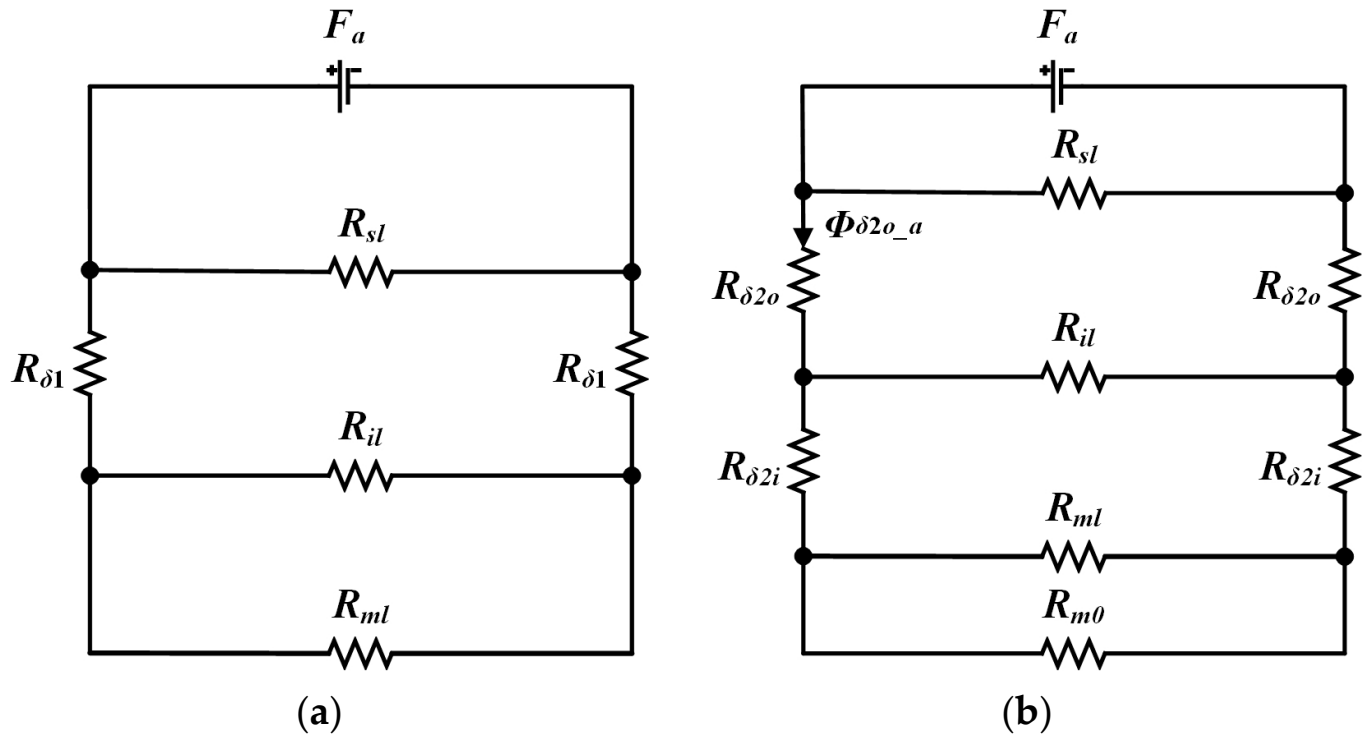


Figure 8. EMC of armature reaction: (a) Area I; (b) Area II.

Based on Figure 8 and expression of the EMC parameters, the flux density under the armature current excitation in the two areas can be calculated by the following equations. If the armature flux in Area II has the same direction as the PM flux, the field would be strengthened and the operation point of the ferromagnetic core would easily rise to the saturation region, which would result in a force reduction.

$$B_{\delta1_a} = \frac{4N_p Ni}{(2R_{\delta1} + R_{ml} \| R_{il}) [\pi(D_{si} - 2\delta_1) + 2N_p \delta_1] (2L_s - L_m + W_g)} \quad (4)$$

$$B_{\delta2o_a} = \frac{4N_p Ni}{\{2R_{\delta2o} + [(R_{ml} \| R_{m0} + 2R_{\delta2i}) \| R_{il}]\} (\pi D_{si} k_{so} + 2N_p \delta_{2o}) (L_m - W_g)} \quad (5)$$

3.3. Derivation of Electromagnetic Force

After preceding the separate analysis of PM and armature fields, the flux density distributed along axial direction (z axis) can be generally depicted as shown in Figure 9. From this figure, it can be seen that due to the difference in air-gap length, Area II has a higher flux density than that in Area I, and when including the armature reaction, the flux density in one side would be strengthened while weakened in the other side. In this case, the mover would be driven towards the side with the higher flux density. Moreover, it can be concluded that if ignoring the saturation effect, the superposition law can be used directly and the flux density in each area is equal to the algebraic sum of the flux density generated by the PMs and armature current.

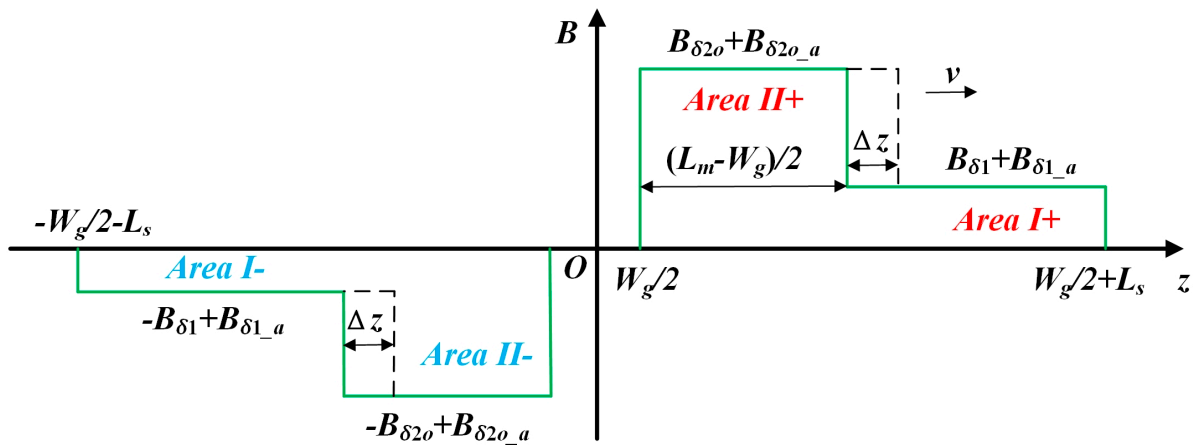


Figure 9. Ideal flux density distribution on the moving direction without considering the saturation effect (z axis) (“+” means flux density being strengthened and “−” means being weakened).

Hence, if ignoring the saturation effect, when the mover is displaced a certain distance (Δz), the flux that links with the armature winding would also change as described in (6). Then, the electromagnetic force can be derived by the virtual displacement theorem, with the expression shown in Equation (7).

$$\Delta\Phi = 2B_{\delta 2o}(\pi D_{si}k_{so} + \delta_{2o})\Delta z - 2B_{\delta 1}[\pi(D_{si} - \delta_1)k_{so} + \delta_1]\Delta z \quad (6)$$

$$F_{em} = i \frac{N\Delta\Phi}{\Delta z} = \{2B_{\delta 2o}(\pi D_{si}k_{so} + \delta_{2o}) - 2B_{\delta 1}[\pi(D_{si} - \delta_1)k_{so} + \delta_1]\}Ni \quad (7)$$

However, because the abovementioned saturation effect occurs in Area II+, actually the flux density cannot reach as high as depicted in Figure 8. When considering the saturation effect of the ferromagnetic core, the flux density in Area II+ would decline somewhat compared with the algebraic sum of $B_{\delta 2o}$ and $B_{\delta 2o_a}$, as shown in Figure 10.

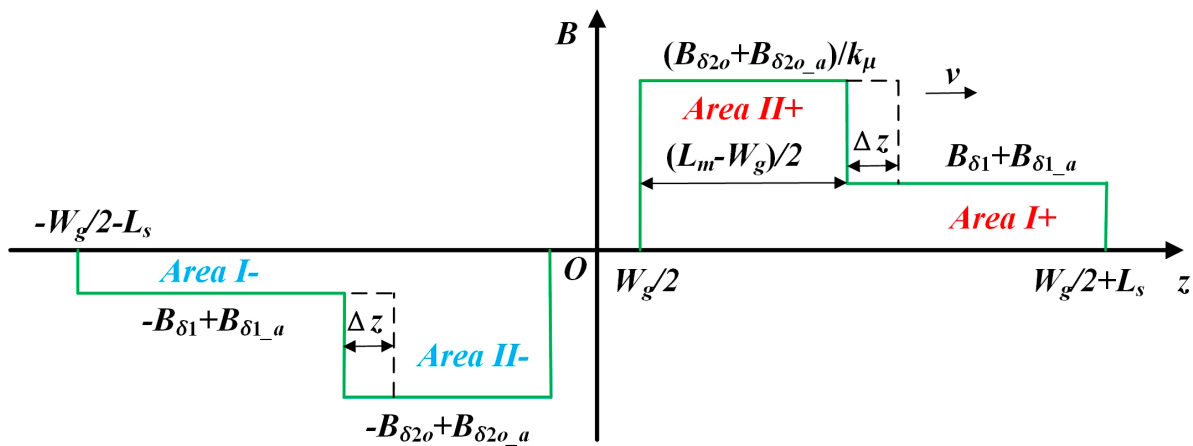


Figure 10. Flux density distribution on the moving direction (z axis) when considering the saturation effect.

Here, a saturation coefficient (k_{μ}) is introduced to quantitatively describe this reduction, which is defined by Equation (8). Hence, to calculate k_{μ} , the saturation value of the flux density ($B_{\delta 2o_sat}$) is required, which needs be determined by the nonlinear EMC of Area II+, as shown in Figure 11.

$$k_{\mu} = \frac{B_{\delta 2o} + B_{\delta 2o_a}}{B_{\delta 2o_sat}} \quad (8)$$

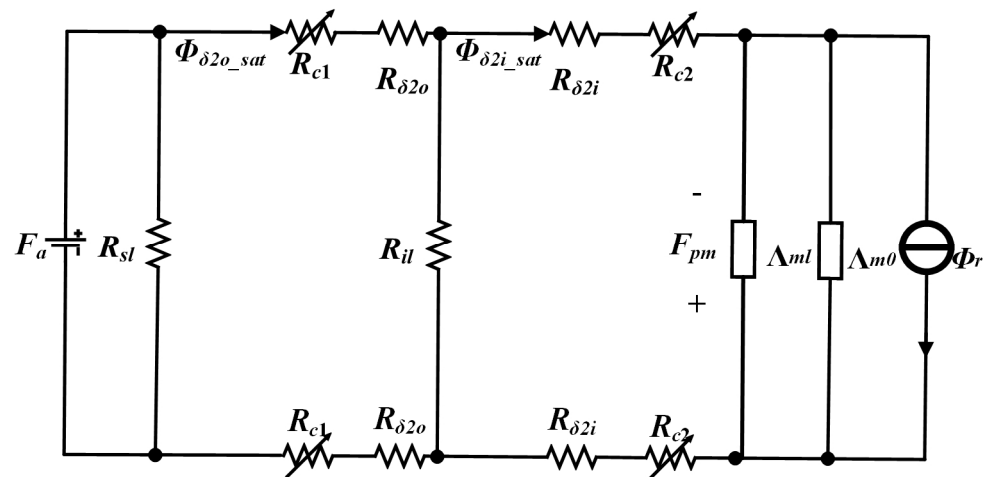


Figure 11. Nonlinear EMC of Area II+.

In this figure, the variable reluctances of the iron cores are taken into consideration, where R_{c1} is the reluctance including the outer stator and mover core, and R_{c2} is related to inner stator core. For their calculation expression, please refer to Appendix A.

The value of these reluctances is determined by the flux density within it, and therefore, an iterative procedure is required to solve them and finally calculate $B_{\delta 2o_sat}$. Figure 12a shows the iterative procedure combined with the nonlinear EMC and fitting BH curve, as depicted in Figure 12b. From Figure 12b, just because the relative permeability of ferromagnetic material varies with the flux density within it, R_{c1} and R_{c2} would also change with $B_{\delta 2o_sat}$ and $B_{\delta 2i_sat}$. So, if the MMF drops on R_{c1} and R_{c2} (F_{Rc1} and F_{Rc2}), calculated by assuming that $\Phi_{\delta 2o_sat}$ cannot satisfy Kirchhoff's Law of nonlinear EMCs, the initial $\Phi_{\delta 2o_sat}$ has to be updated until it meets the convergence condition.

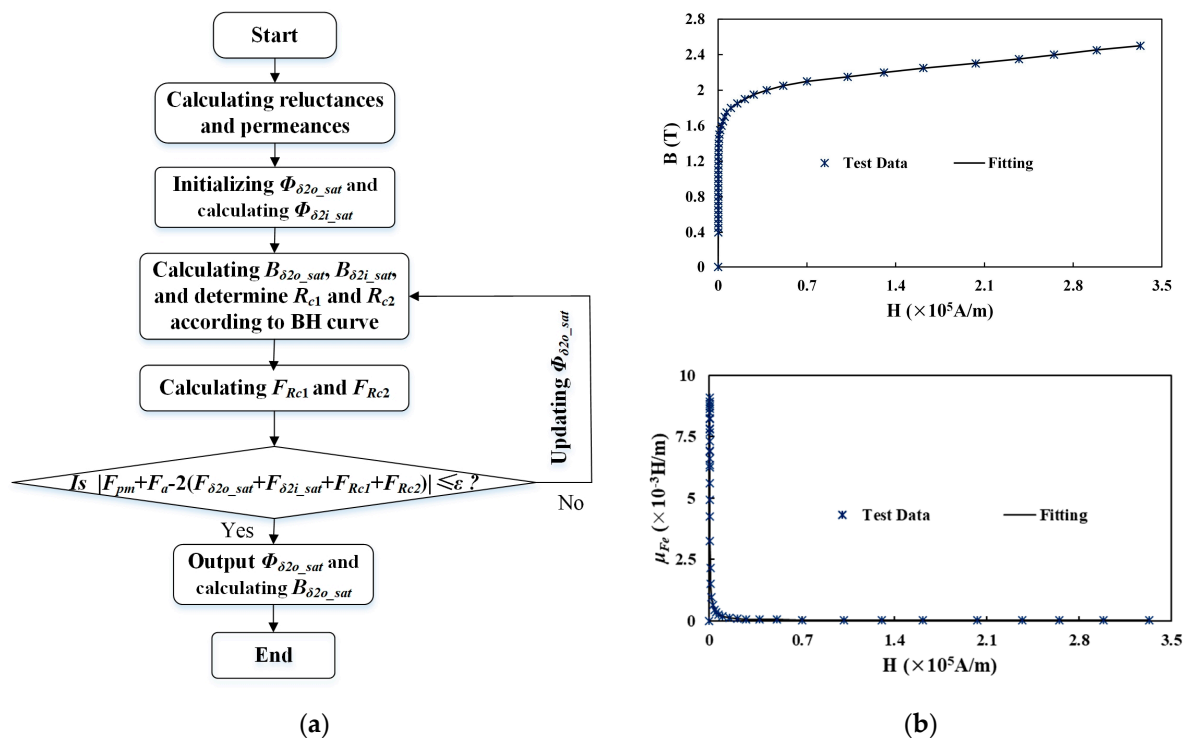


Figure 12. Calculation procedure for $B_{\delta 2o_sat}$: (a) iterative calculation flow chart; (b) fitting the B–H curve and permeability of the ferromagnetic material.

After the iterative procedure, k_μ can be obtained and the expression of electromagnetic force considering the saturation effect can be rewritten as

$$F_{em_sat} = \left\{ \left(1 + \frac{1}{k_\mu}\right) B_{\delta 2o} (\pi D_{si} k_{so} + \delta_{2o}) - 2 B_{\delta 1} [\pi (D_{si} - \delta_1) k_{so} + \delta_1] \right\} Ni \quad (9)$$

Because of the saturation effect, $B_{\delta 2o_sat} < B_{\delta 2o} + B_{\delta 2o_a}$, so $1/k_\mu < 1$. Therefore, the modified $F_{em_sat} < F_{em}$ has properly described quantitatively the force reduction in this case.

4. Main Size Selection and Performance Comparison

To further enhance the performance of the ISMTLOM and make a fair comparison with the CSMTLOM described in [27], its size parameters have to be optimized under some restrictions of the same outer diameter and copper loss. The selected size parameters include the split ratio (k_s), slot open ratio (k_{so}), and mover versus stator pole ratio (k_{ms}). When the former restriction is satisfied, these parameters would affect the cross area of the coil and the air-gap flux. Namely, they would significantly influence both the electric and magnetic loads, which can finally determine the output force [28]. Moreover, because the PMs of the ISMTLOM are only placed on the inner stators, its mover would suffer from the centripetal attractive force from the inner PMs, even when the winding is an open circuit. Hence, the inner/outer air-gap ratio ($k_\delta = \delta_{2i}/\delta_{2o}$) parameter is required to be introduced and investigated, to assure the reasonable centripetal attractive force for safe operation.

4.1. Selection of Split Ratio (k_s) and Slot Open Ratio (k_{so})

These two parameters have significant influence on the slot area (A_s) when D_{so} is fixed, as illustrated by Equation (10). And then they would consequently impact the winding current density (J) if the copper loss remains the same as in the CSMTLOM, as described in Equation (11), where P_{cl} is the copper loss, ρ_{cu} is the resistivity of copper, k_f the fill factor of slot, and L_{end} the length of coil end, which can be evaluated by Equation (12).

$$A_s = \frac{\pi}{N_p} \left[\frac{(1 - k_s k_{so})^2 - k_s^2}{4} - \frac{k_{so} k_s (1 - k_s - k_s k_{so})}{2} \right] D_{so}^2 \quad (10)$$

$$J = \sqrt{\frac{2P_{cl}}{N_p \rho_{cu} k_f A_s [2(2L_s + W_g) + L_{end}]}} \quad (11)$$

$$L_{end} = \frac{2\pi k_s k_{so} D_{so}}{N_p} + \frac{\pi N_p k_f A_s}{[N_p (1 - k_s) - \pi k_s k_{so}] D_{so}} \quad (12)$$

As depicted in Figure 13, changing k_s and k_{so} would result in a variation in the current density (J) under the abovementioned constraints. It can be found that the output force does not always increase with J . From Figure 13a, an optimal choice of (k_s , k_{so}) for the ISMTLOM is (0.35, 0.6); for the CSMTLOM, according to [24], the optimal value for these two parameters is (0.45, 0.6).

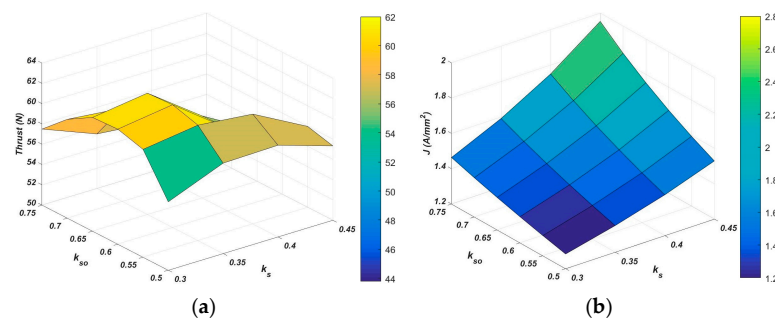


Figure 13. Influence of k_s and k_{so} on output thrust: (a) variation of thrust; (b) variation of J .

4.2. Selection of Mover/Stator Pole Ratio (k_{ms}) and Outer/Inner Air–Gap Length (k_δ)

The value of k_{ms} can affect the effective area of the air gap, and consequently influence the output force. Generally, an increase in k_{ms} would lead to an increase in the air–gap flux. However, it should be noticed that the larger the k_{ms} is, the closer the mover poles are, and thus there is a higher flux leakage between each two adjacent mover cores. It means that an excessive k_{ms} would not lift the thrust but would instead decrease. As shown in Figure 14, obviously, the optimal value of k_{ms} should be around 1.3. But taking the mover weight into consideration, the increment of thrust from $k_{ms} = 1.2$ to 1.3 is not that notable, whilst it increases linearly with k_{ms} . Thus, to obtain a lighter mover, a more reasonable choice for k_{ms} should be 1.2. Also, according to previous research results in [24], the optimal value of k_{ms} for a CSMTLOM is 1.1.

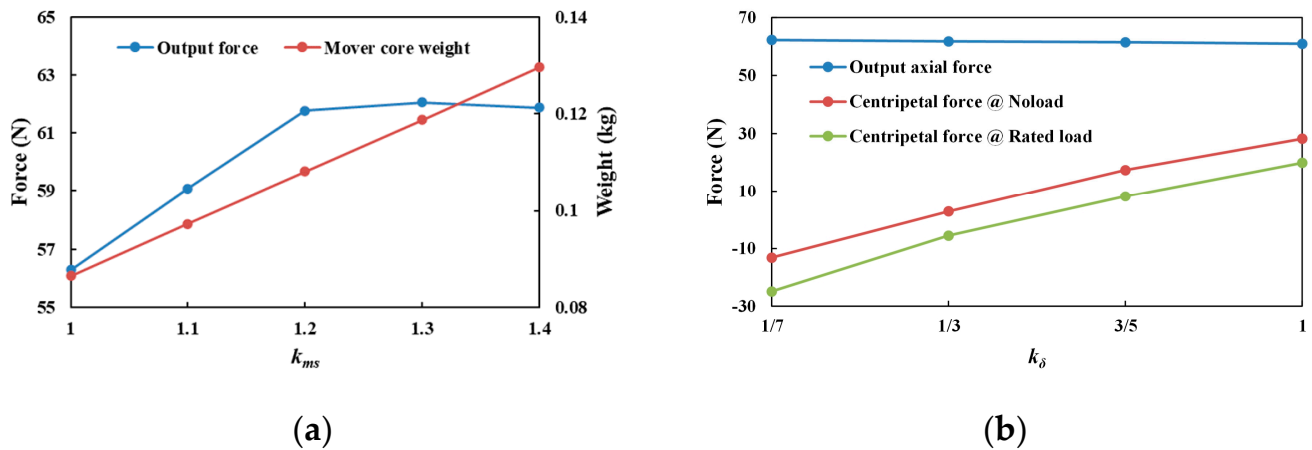


Figure 14. Selection of k_{ms} and k_δ : (a) output force and mover core weight versus k_{ms} ; (b) output force and axial force versus k_δ .

While the mover vibrates between the inner and outer stators, besides being subjected by the axial electromagnetic force, it also suffers from the centripetal attractive force, which is related to the ratio of the outer/inner air–gap length, k_δ . As shown in Figure 14b, while the total air–gap length remains at 0.8 mm to meet the F_{em} requirement, the centripetal attractive force would increase with increasing k_δ , regardless no–load or rated condition. The negative value means that the centripetal attractive force is directed toward the outer stator, while it is opposite for a positive value. To minimize the F_a , it is easy to conclude from Figure 14b that $k_\delta = 1/3$ is an optimal choice. Furthermore, it can be seen from this figure that the output force does not decrease with increasing k_δ , which means that changing k_δ to minimize F_a has almost no impact on the output force. Thus, in general, $k_\delta = 1/3$ is the optimal value, which indicates $\delta_{2i} = 0.6$ mm and $\delta_{2o} = 0.2$ mm.

4.3. Electromagnetic Performance Analysis and Comparison with the CSMTLOM

In this part, the 3D FEA method (carried out by ANSYS Electromagnetics Suite version 2021R1) is employed to calculate the important performance indices, including the flux distribution, back EMF, static and dynamic electromagnetic force, etc. The main specifications of the CSMTLOM and ISMTLOM are listed Table 1.

Figure 15 shows the mesh of the 1/6 3D FEA model and the flux density distribution at the balance position within a wide load condition. While the mover rests at the balance point, for both the CSMTLOM and ISMTLOM, it is found that the flux distribution in each side of the stator is symmetrical under the no–load condition, while asymmetrical under the rated condition. Therefore, the mover would be driven toward the flux–strengthened side, which is coincidence with the former analysis. Moreover, it can be seen from Figure 15b that the flux density in the flux–strengthen side of the stator is much higher than the

weakened side, which is the so-called axial local saturation effect. It would become severe under the overload condition and significantly influence the output force.

Table 1. Main specifications of the CSMTLOM and ISMTLOM.

Items	Symbol	CSMTLOM	ISMTLOM	Unit
Copper loss	P_{cl}	13	13	W
Stator outer diameter	D_{so}	148	148	mm
Stator stacking length	L_s	32	30	mm
Mover stacking length	L_m	32	32	mm
Stator interval length	W_g	6	6	mm
Total mechanical air gap	—	0.4	0.8	mm
Coil turns	N	350	350	turns
Split ratio	k_s	0.45	0.35	—
Slot open ratio	k_{so}	0.6	0.6	—
Mover/stator pole ratio	k_{ms}	1.1	1.2	—
Outer/inner air-gap ratio	k_δ	—	1/3	—
Pure copper filling factor	k_f	0.45	0.45	—
Width of PM	W_{pm}	13	13	mm
Thickness of PM	T_{pm}	3	3	mm

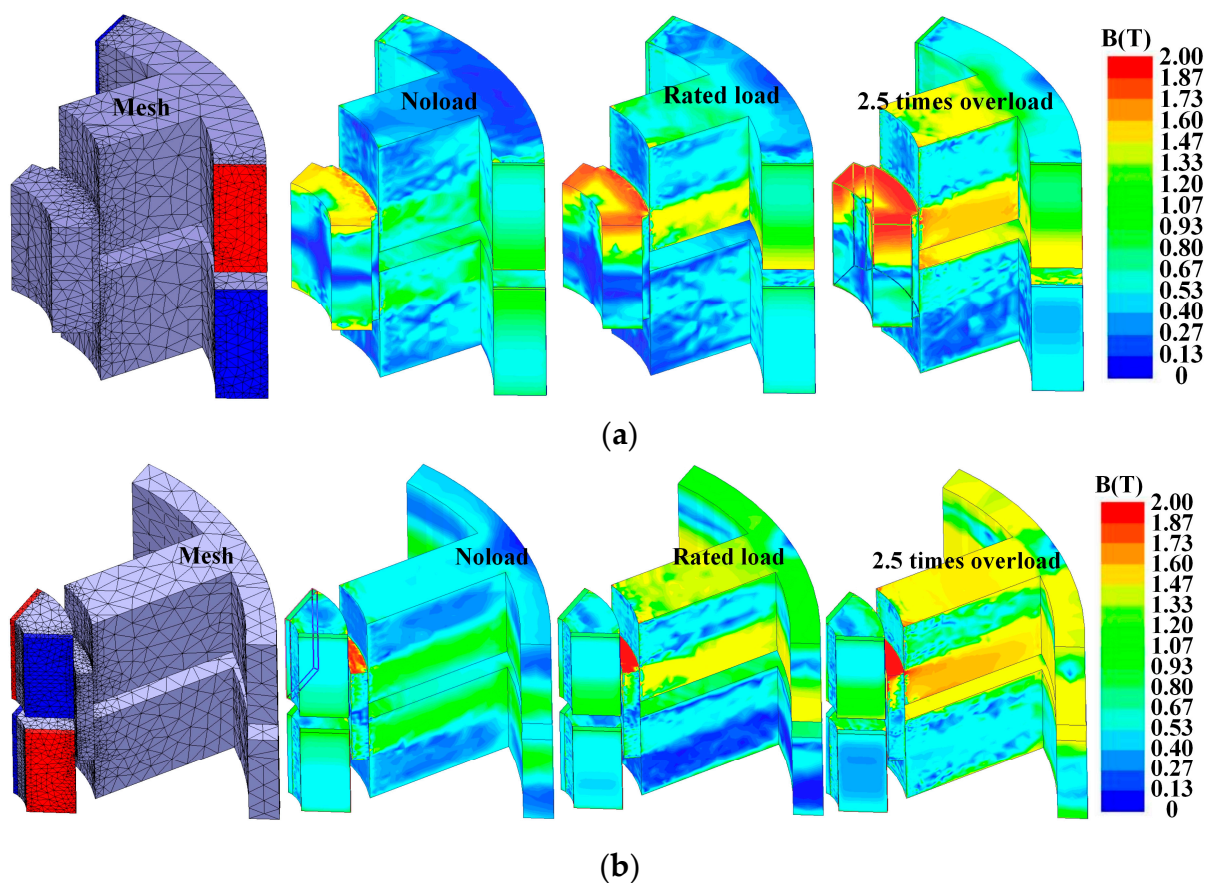


Figure 15. Flux density: (a) under the different load conditions of CSMTLOM; (b) under the different load conditions of ISMTLOM.

Based on Figure 15, the axial air-gap flux density under no load and rated load are depicted in Figure 16. From this figure, the results from the FEA and the established EMC are compared, which validate the effectiveness of the analytical model. The maximum analytical calculation error comparing with 3D FEA in average flux density is 5.00%.

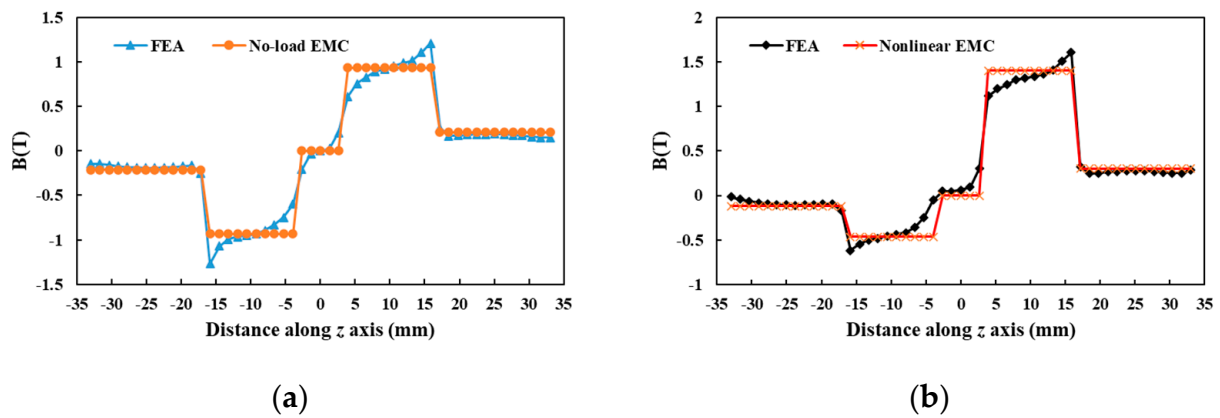


Figure 16. Air-gap flux density of the ISMTLOM along the z axis: (a) no load; (b) loaded.

Next, the static simulation is carried out when the winding is fed by a dc current, which aims to obtain the static axial force on the mover. Figure 17 shows the axial force versus mover displacement for both the ISMTLOM and CSMTLOM, which contains three curves, i.e., total force, detent force, and electromagnetic force. From this figure, it can be found that the detent force of the ISMTLOM is nearly zero within the designed stroke (± 10 mm), and is also lower than that of the CSMTLOM. Moreover, when subtracting the detent force from total force, the pure electromagnetic force generated by the current can be obtained. It can be determined that the average of the electromagnetic force of these two LOMs are similar (61.92 N for the ISMTLOM and 57.48 N for the CSMTLOM), but the waveform of the electromagnetic force of the ISMTLOM is much smoother than that of the CSMTLOM, which indicates less force ripple during the operation.

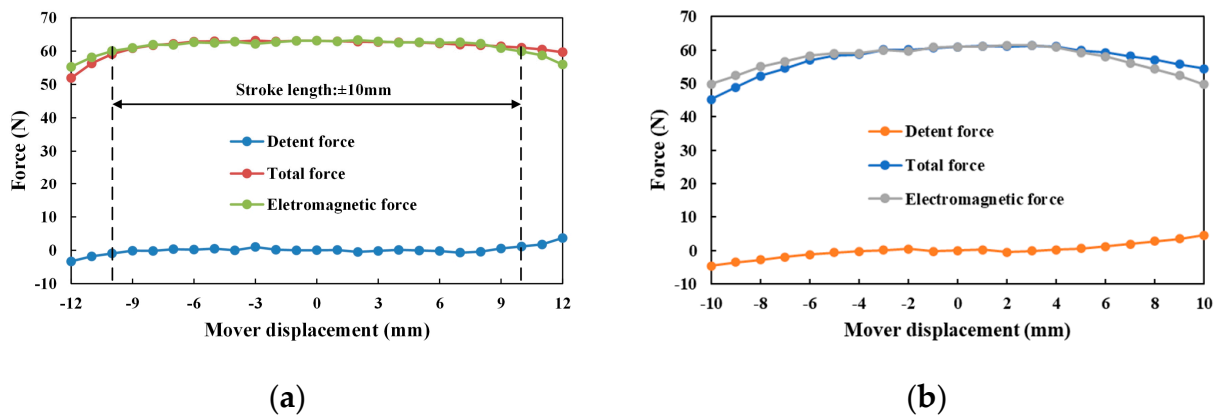


Figure 17. Static force: (a) ISMTLOM; (b) CSMTLOM.

Also, the curve of force versus current can also be obtained, as shown in Figure 18, demonstrating the overload capability of the proposed LOM. From Figure 18a, it can be seen that the electromagnetic force increases linearly with the current when the current is smaller than the rated value (p.u. = 1). However, if the current continues to increase after reaching the rated load, the force would no longer increase linearly because the saturation occurs in the iron core. Therefore, from Figure 18a, one can easily know that under the overload condition, Equation (7) cannot accurately calculate the electromagnetic force. However, benefiting from the introduction of k_{μ} to involve the saturation effect, the modified Equation (9) can still attain better agreement with the FEA results, of which maximum relative calculation error is only 3.5% during the range from no load to 2.5 times overload. Moreover, from Figure 18b, it can also be shown that when comparing with the CSMTLOM, the proposed ISMTLOM has better overload capability because its force versus current characteristic is more linear than that of the CSMTLOM. This mainly results

from the lighter saturation of the ISMTLOM in a wide load range, and the maximum force improvement at 2.5 times over load is nearly 8%.

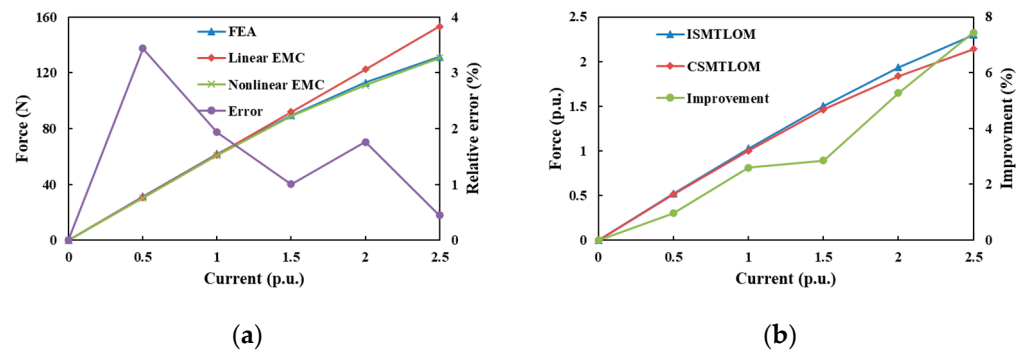


Figure 18. Overload capability: (a) verification of EMC results; (b) comparison with the CSMTLOM.

Consequently, the dynamic performance is also analyzed, including the back EMF and axial driving force generated by sinusoidal current. Figure 19 shows the back EMF and its FFT spectrum when the mover is vibrating at a sinusoidal speed. It can be seen that the analytical method result is identical with that from the 3D FEA, and the amplitude of the back EMF is 150.53 V, whose total harmonic distortion (THD) is only 1.10%.

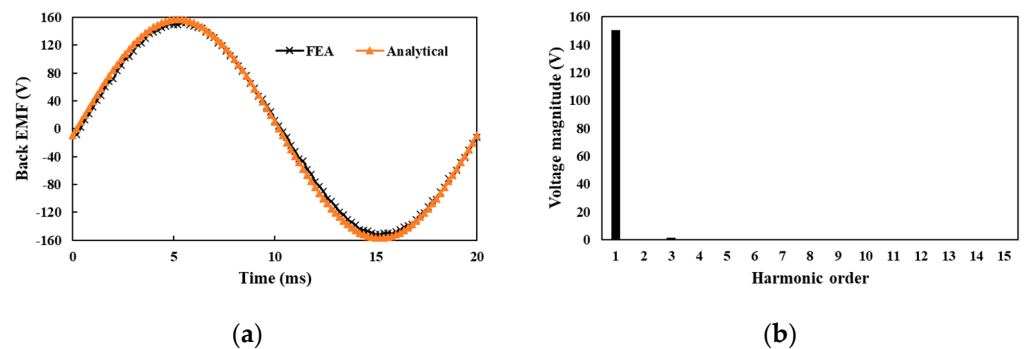


Figure 19. Back EMF: (a) waveform; (b) FFT spectrum.

Therefore, benefiting from the high-quality back EMF, the dynamic electromagnetic force is also a good sinusoidal wave, as shown in Figure 20, in which the THD is also only 1.10%, and the better the sine of the force, the lower the high-frequency vibration.

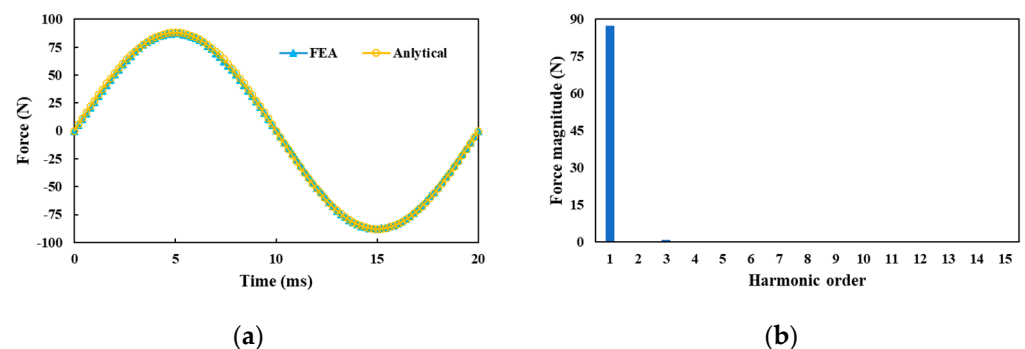


Figure 20. Dynamic force: (a) force versus time under sinusoidal supply; (b) FFT spectrum.

The overall comparison between the analytical method and the 3D FEA is summarized in Table 2, together with the calculation time. Generally speaking, the proposed analytical model with an introduced saturation coefficient can reach a relatively high accuracy, and can significantly reduce the force calculation time.

Table 2. Comparison of the ISMTLOM results from the 3D FEA and the established analytical model.

Items	3D FEA	Analytical	Error
Rated static force within stroke	61.92 N	61.25	−1.10%
2.5 times over load force	131.28 N	130.69 N	−0.45%
Amplitude of back EMF	152.12 V	157.45 V	+3.50%
Amplitude of dynamic force	87.28 N	88.03 N	+1.16%
Static performance calculation time	10 min	≤0.5 s	—

As for LOMs, when applied to drive a reciprocating device, the spring is necessary to form a harmonic vibration system to save energy. Only the driving force has the same frequency as its natural force; such a system can reach the maximum stroke by the least current. The eigen frequency of this harmonic vibration system is determined by

$$f = \frac{1}{2\pi} \sqrt{\frac{k}{m}} \quad (13)$$

where f is the driving frequency, k is the required spring stiffness, and m is the mover weight. So, the lighter the mover core, the less spring stiffness is required. Comparing the CSMTLOM, one more advantage of the ISMTLOM is that the mover core is much thinner, and thus the mover core is much lighter. So, as listed in Table 3, the mover core weight of the ISMTLOM is only about 1/5 of the CSMTLOM, and the required spring stiffness can also be greatly reduced, which can lower the cost and be easy to assemble.

Table 3. Mover weight and least required spring stiffness for the CSMTLOM and ISMTLOM.

Items	Symbol	CSMTLOM	ISMTLOM	Unit
Mover core weight	m	0.48	0.1	kg
Required spring stiffness	k	≥47.4	≥9.9	kN/m

4.4. Performance Comparison under Open and Closed–Loop Control System

To investigate and compare the overall system performance of these two PMLOMs, the model of the two machines was built based on (14), according to Figure 1b. The most important parameters of these two machines are listed in Table 4.

$$\begin{cases} u(t) = Ri(t) + L \frac{di(t)}{dt} + k_i \frac{dz(t)}{dt} \\ m \frac{d^2z(t)}{dt^2} + c \frac{dz(t)}{dt} + kz(t) = F(t) = k_i i(t) \end{cases} \quad (14)$$

Table 4. Electromagnetic parameters in the model of the CSMTLOM and ISMTLOM.

Parameters	Symbol	CSMTLOM	ISMTLOM	Unit
Resistance	R	12.3	7.86	Ω
Inductance	L	0.86	0.60	H
Force constant	k_i	55.18	48.16	N/A
Damping coefficient	c	25	25	N•s/m

It can be seen from this table that the resistance and inductance of the ISMTLOM are lower than that of the CSMTLOM, because of the larger slot area and longer air–gap length, respectively. Hence, to keep their copper loss the same, the ISMTLOM current should be larger than that of the CSMTLOM, resulting in a lower force constant.

According to (14), the equivalent circuit of these two LOMs can be described as shown in Figure 21a, which can be separated into the electrical and mechanical parts, coupled through the current and the mover speed. From the mechanical part, it can be shown that, concerning the PMLOMs, the mechanical load can be regarded as the damping coefficient,

while the mass and mechanical springs only work as the energy storage components. Afterward, the displacement closed-loop control system of these two LOMs is built in Matlab/Simulink, with the diagram shown in Figure 21b, adopting an H-bridge converter, PI controller, and SPWM modulation. In this figure, z_m^* and z_m represent the given displacement amplitude and the actual amplitude, and u_M is the modulation ratio. The optimal parameters for the PI controller are 20 and 1000 for k_p and k_I .

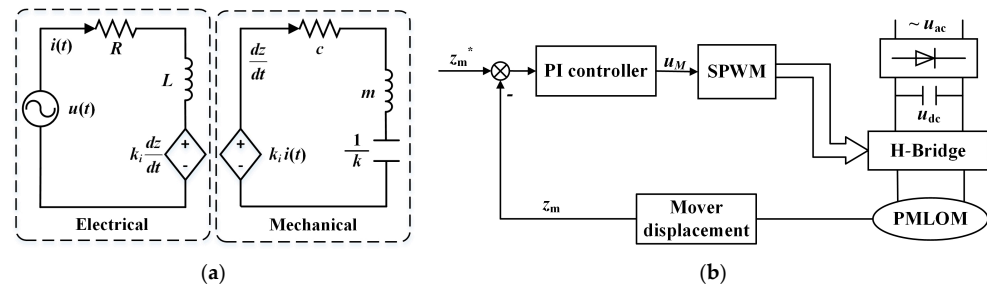


Figure 21. Diagram of overall closed-loop control system of the LOM: (a) equivalent circuit; (b) control diagram.

Based on this control system, the performance of these two PMLOMs is investigated and compared under the same damping coefficient and different stroke. As shown in Figure 22, the curves of voltage and current amplitude versus driving frequency are depicted. It can be found that as to these two PMLOMs, regardless the stroke, when the driving frequency is equal to the resonance frequency of 50 Hz, the current converges to its minimum value. It indicates that both can utilize the mechanical resonance to reduce the driving current for saving energy.

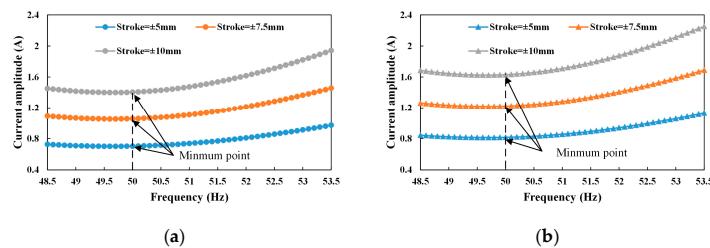


Figure 22. Current amplitude vs. frequency under the same load: (a) CSMTLOM; (b) ISMTLOM.

Moreover, the power factor is also compared in Figure 23a. From this figure, it can be shown that the power factor of the ISMTLOM is slightly larger than that of the CSMTLOM because of its lower inductance. Furthermore, benefiting from its lighter moving mass, its power factor is stable within a wide frequency band. It is because the moving mass works like an inductance in the mechanical system, as shown in Figure 21a, leading to a drastic change during the frequency variation.

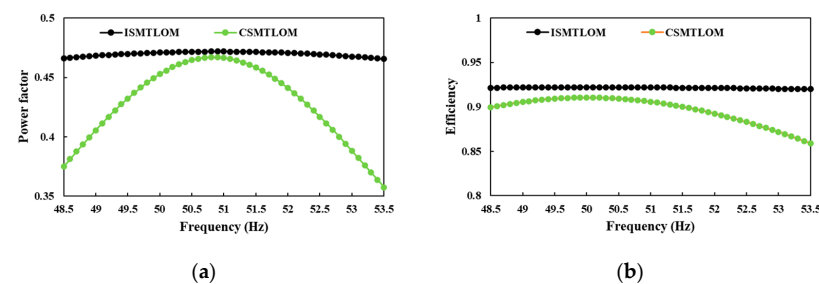


Figure 23. Voltage and current amplitude vs. frequency under the ISMTLOM rated load: (a) voltage amplitude; (b) current amplitude.

Consequently, the efficiency is also calculated via (15), where z_m , U_m , and I_m are the amplitudes of displacement, voltage, and current, respectively; θ_{i-z} is the phase between current and displacement; and $\cos \varphi$ is the power factor.

$$\eta = \frac{2\pi f z_m k_i I_m \cos(\theta_{i-z} - \frac{\pi}{2})}{U_m I_m \cos \varphi} \quad (15)$$

The efficiency versus frequency curve is depicted in Figure 23b. It can be seen that the efficiency of ISMTLOM reaches its maximum under the resonance frequency, and the maximum efficiency of ISMTLOM is slightly larger than that of CSMTLOM because of the higher rated force. Moreover, the efficiency curve of ISMTLOM exhibits more stability than that of CSMTLOM, owing to its more stable power factor.

In addition to the characteristics noted above, the dynamic response for these two PMLOMs is also compared. As shown in Figure 24, the curves of the voltage, current, and displacement from the starting to steady state are depicted, under the same load, resonance frequency, and stroke length. It can be seen that for both of these PMLOMs, their adjusting time, 0.254 s, is nearly the same from starting to steady state, and the phase between the current and displacement is 90 degrees, which is identical to the mechanical resonance. Worth noting is that the required voltage of the ISMTLOM is lower than that of the CSMTLOM, even the former has the larger operation current. It is mainly because the inductance of the ISMTLOM is lower than the CSMTLOM.

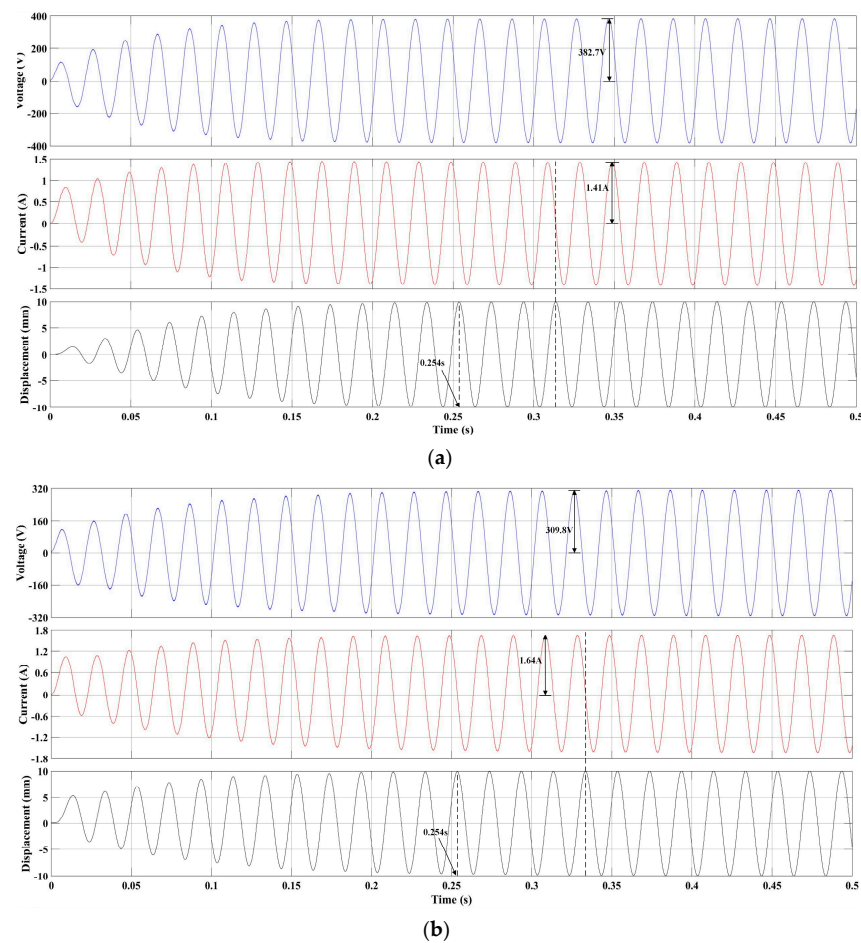


Figure 24. Dynamic response when starting under the given step displacement of ± 10 mm: (a) CSMTLOM; (b) ISMTLOM.

Therefore, when reaching a similar dynamic response under the same working condition, owing to the reduction in the moving mass and the inductance, the proposed

ISMTLOM can obtain a more stable power factor and efficiency within a wide frequency band, and a lower requirement on input voltage.

5. Conclusions

In this paper, a novel ISMTLOM is proposed, and the corresponding segmental analytical model considering the local saturation effect is built for a rapid performance calculation. Under the similar volume and copper loss restriction, the most important size parameters are optimally selected and, consequently, the quantitative comparisons have also been made with the CSMTLOM. Sufficient simulation results have demonstrated that, whilst reaching the same output force, the ISMTLOM provides several advantages over the CSMTLOM, i.e., lower detent force, smoother force curve during the stroke, better overload capability, and less demand for spring stiffness. Specific findings are described as follows:

- (1) Under the same size and copper loss limitation, the ISMTLOM can reach 61.92 N static force, in which the curve is smoother and the amplitude is slightly larger than the CSMTLOM.
- (2) The overload force of the ISMTLOM is larger than that of the CSMTLOM due to the lower saturation degree, reaching nearly 8% improvement under a 2.5 times overload condition.
- (3) Benefiting from the separate winding and PMs on the out and inner stators, respectively, the yokeless mover core is redesigned, reducing its weight to 0.1 kg, and as a result, the required spring stiffness could be reduced by at least 33.5 kN/m.
- (4) With the introduction of the saturation coefficient, the static force calculation under a wide load range is rapid and accurately calculated, with a maximum error of less than 3.5%, and the time consumption is reduced from 10 min to 0.5 s.
- (5) Owing to the reduction in the moving mass and the inductance, the proposed ISMTLOM can obtain a more stable power factor and efficiency within a wide frequency band, and a lower requirement on input voltage.

Further experimental study would show how well the ISMTLOM works in a practical linear reciprocating drive system.

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Appendix A

The calculation expressions for the EMC parameters are listed in Table A1. Note that for both the CSMTLOM and ISMTLOM, it is better to maintain the same flux density in the stator as that in the teeth, because the flux flow through them is nearly the same. Hence, the thickness of the yoke should be determined by Equation (A1).

$$T_{sy} = \frac{\pi k_s k_{so} D_{so}}{2N_p} \quad (A1)$$

Table A1. Calculation expressions for the EMC parameters.

Item	Calculation Expression
A_{m0}	$A_{m0} = \frac{\mu_0 \mu_r W_{pm} (L_m - W_g)}{2T_{pm}}$
A_{ml}	$A_{ml} = \frac{\mu_0 (L_m - W_g)}{2\pi} \left[\ln \frac{\pi (k_s D_{so} - 2\delta_1 - 2T_{pm})}{N_p T_m} + \ln \frac{\pi (k_s D_{so} - 2\delta_1) (1 - k_{so} k_{ms})}{N_p T_m} \right] + \frac{\mu_0 W_{pm}}{\pi} \ln \frac{\pi (k_s D_{so} - 2\delta_1 - T_m)}{N_p T_m}$
$R_{\delta 2i}$	$R_{\delta 2i} = \frac{4\delta_{2i}}{\mu_0 (\pi k_s D_{so} k_{so} + 2N_p \delta_{2i}) (L_m - W_g)}$
$R_{\delta 2o}$	$R_{\delta 2o} = \frac{4\delta_{2o}}{\mu_0 (\pi k_s D_{so} k_{so} + 2N_p \delta_{2o}) (L_m - W_g)}$
$R_{\delta 1}$	$R_{\delta 1} = \frac{4N_p \delta_1}{\mu_0 (\pi k_s k_{so} D_{so} + N_p \delta_1) (2L_s - L_m + W_g)}$
A_{il}	$A_{il} = \frac{N_p \mu_0 (L_m - W_g)}{2\pi (1 - k_{so} k_{ms})} \ln \frac{k_s D_{so} - 2\delta_{2o}}{k_s D_{so} - 2\delta_{2o} - 2T_m}$
A_{sl}	$A_{sl} = \frac{\mu_0 N_p L_s}{2\pi} \ln \frac{N_p - (\pi + N_p) k_s k_{so}}{N_p k_s (1 - k_{so})}$
R_{c1}	$R_{c1} = \frac{(4\pi N_p D_{so} - 2\pi^2 k_s k_{so} D_{so}) + 4N_p^2 (1 - k_s) D_{so} - 2\pi N_p k_s k_{so} D_{so} + 8N_p^2 T_m}{\mu_{Fe1} \pi N_p k_s k_{so} D_{so} (L_m - W_g)}$
R_{c2}	$R_{c2} = \frac{\pi (k_s D_{so} - 2\delta_1 - W_{pm}) - N_p T_{pm}}{\mu_{Fe2} N_p W_{pm}}$

* μ_{Fe1} and μ_{Fe2} are the permeability of ferromagnetic material, which can be determined from Figure 12b during the iteration.

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