Jurnal
Teknologi

Full Paper

Coupling-Based Turns Distribution for Planar Spiral Coil Antennas in Near Field Magnetic Induction Links

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Article history
Received
31 March 2015
Received in revised form
22 July 2015
Accepted
28 August 2015

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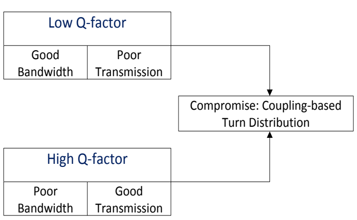
Graphical abstract



Abstract

This paper describes a numerical study of a coupling-based turn-distribution approach to achieving good transmission performance in near field magnetic inductive links using low Q-factor coil antennas. Coil antenna turns are distributed to match an upper reference coupling level, leading to a stronger axial H-field, with a low margin of Q-factor increment from a baseline minimum. Numerical results demonstrate improved transmission efficiency performance relative to a minimal Q-factor coil antenna when the modified antenna is employed in a symmetric bi-directional inductive link. This approach to increasing transmission efficiency indicates prospects for extending the utility of low Q-factor HF-RFID interrogator antennas to include wireless power delivery applications.

Keywords: Coupling, H-field, HF-RFID, Q-factor, coil antenna, magnetic induction

Abstrak

Kertas kerja ini membincangkan kajian berangka ke atas pendekatan asas salingan pada agihan gelung bagi penghantaran yang baik dalam medan magnetic dekat menggunakan antenna gelung dengan factor Q yang rendah. Gelung antenna berkenaan direkabentuk agar sepadan dengan aras rujukan salingan yang membolehkan pencapaian medan magnetik yang lebih tinggi, juga perubahan factor Q yang lebih rendah. Keputusan kajian berangka yang dilakukan menunjukkan peningkatan kecekapan relative kepada factor Q minimum pada antenna yang diubahsuai dan diuji dalam hubungan simetri dua hala. Peningkatan tahap kecekapan melalui pendekatan ini menunjukkan ianya boleh digunakan bagi antenna HF-RFID dengan factor Q yang rendah bagi aplikasi penghantaran kuasa tanpa wayar.

Kata kunci: Coupling, H-field, HF-RFID, Q-factor, coil antenna, magnetic induction

1.0 INTRODUCTION

Near-field magnetic induction has found application in such areas as biotelemetry, item-level tracking, access-control, and e-transactions, through the enabling technologies of High-frequency Radio Frequency Identification (HF-RFID) and Near Field Communications (NFC) at 13.56 MHz. Magnetic induction schemes operate through the linkage of magnetic flux from the near-field of a coil antenna source. These flux linkages are more intense when the distance z between interacting antenna terminals and the wavelength λ of the oscillating field are related as $z << \lambda/2\pi$. The effective range of these links is typically short since inductive systems are quite inefficient when the operating range is greater than the characteristic sizes of the coupled antennas [1,2].

An intuitive approach to enhancing the operational range of a magnetic induction link would be to minimize losses within the coil antenna structure, through high Q-factor designs. However, this is inimical to meeting bandwidth requirements for reliable data transfers, as a consequence of the inverse relationship between Q-factor and bandwidth. Hence, HF-RFID/NFC links usually do not place a premium on high Q-factor coil antennas. In particular [3,4], suggest the use of single-turn structures as interrogator coil antennas in HF-RFID systems in order to realize the lowest possible unloaded Q-factors, and hence the best bandwidth performance. This, however, limits the strength of the interrogator H-field, leading to even shorter operational ranges.

Distribution of coil antenna turns has been studied to increase the operating range of HF-RFID interrogators through H-field enhancement without compromise to minimal unloaded Q-factor levels [5,6]. However, minimal unloaded Q-factors may inhibit link performance if these same coil antennas are required in symmetric bi-directional inductive coupling links, as could occur in power-focused non-data-centric peer-to-peer use-cases. The associated circuitries at paired terminals present additional loading effects, resulting in further reduction of Q-factors, increased losses, and hence decreased efficiency. Consequently, it is necessary to investigate geometric approaches to realizing low Q-factor coil antennas, while still achieving appreciable levels of energy transmission efficiency in symmetric bi-directional links.

This paper presents a numerical study of a coupling-based approach to achieving good transmission performance with low Q-factor coil antennas employed in symmetric bi-directional links. In the proposed approach described in Section 2, the turns of a coupled pair of planar square multi-turn coil antennas are spaced to reduce their Q-factor levels, while matching the coupling level that can be achieved using a coupled pair of conventional coil antennas of the same number of turns. Numerical results in Section 3 show that it is possible for a spaced multi-turn coil antenna to achieve a reduced Q-factor level and stronger magnetic field compared to the conventional multi-turn design. In addition, the achieved coupling performance allows the realization of better transmission efficiencies relative to a minimal Q-factor single-turn design. The paper is concluded in Section 4.

2.0 COIL ANTENNA DESIGN METHODOLOGY

The potential for a greater interrogation range in an HF-RFID link is evident in the ability of the interrogator antenna to excite a stronger voltage in a tag coil. The induced voltage is modelled by Faraday's law:

$$V_{ind} = j\omega\mu_0 H N_{Rx} A_{Rx} \tag{1}$$

For a tag coil antenna with a predetermined number of turns N_{RX}, and a predetermined area A_{RX}, this induced voltage can be augmented by increasing the magnetic field strength H from the interrogator coil antenna, where μ_0 is the vacuum permeability.

Biot-Savart's law from classical magnetostatics allows the derivation of the H-field at a point z along the z-axis of an N-turn interrogator coil antenna in the x-y plane as (see Appendix):

$$H = \frac{NI}{2\pi\left(\frac{z^2}{l^2}+\frac{1}{4}\right)\sqrt{z^2+\frac{l^2}{2}}}, \tag{2}$$

where the current I flows through the planar square interrogator coil antenna, whose sides have a length l. With fixed current levels and antenna dimensions, H increases with the number of turns N. If this same coil antenna is used in a bi-directional inductive coupling link, the magnetic flux at the receive coil antenna as a consequence of the transmit coil H-field H_{RX} is given by

$$\Psi_{Tx,Rx} = N_{Rx} A_{Rx} \mu_0 H_{Tx} \tag{3}$$

In this bi-directional link, the mutual inductance that develops between the transmit and receive coil antennas is given by the relation

$$M_{Tx,Rx} = \Psi_{Tx,Rx} / I \tag{4}$$

Also, the coupling between the transmit and receive coil antennas can be expressed as a coupling coefficient:

$$k_{Tx,Rx} = M_{Tx,Rx} \Big/ \sqrt{L_{Tx} L_{Rx}}, \tag{5}$$

where L_{Tx} and L_{Rx} are the external inductances of the transmit and receive coil antennas respectively. Meanwhile, the unloaded Q-factors of these antennas can be derived as

$$Q_{Tx,Rx} = \omega L_{Tx,Rx} / R_{Tx,Rx}, \tag{6}$$

where $R_{Tx,Rx}$ refer to their respective resistances at the desired operating frequency ω. In a symmetric bi-directional inductive coupling link, the coupled coil antennas will be similar, hence, $L_{Tx} = L_{Rx}$ and $R_{Tx} = R_{Rx}$.

Equation (5) shows that the coupling level between multi-turn symmetric coil antennas can be enhanced by reducing the inductances of the antennas. Similarly, equation (6) reveals these reduced inductances lead to lower unloaded Q-factor levels for the multi-turn coil antennas.

The inductance of an N-turn planar square coil antennas is modelled as[7]

$$L_{self} = 0.635\mu N^2 l_{avg}\left[\ln\left(\frac{2.07}{\vartheta}\right)+0.18\vartheta+0.13\vartheta^2\right], \tag{7}$$

where the antenna fill factor ϑ and average side-length l_{avg} are defined as

$$\vartheta = \frac{l_{out}-l_{in}}{l_{out}+l_{in}}, \qquad l_{avg} = \frac{l_{out}+l_{in}}{2} \tag{8}$$

For uniform turn spacing s and width w, the length of the i-th turn l_i can be derived from the length of the innermost turn l_{in} as

$$l_i = l_{in} + 2i(w+s) - 2s, \qquad i = 1,\ldots,\mathrm{N}, \tag{9}$$

so that the length of the outermost turn $l_{out} = l_N$. The mutual inductances existing between the i-th and j-th turns on the planar square coil antenna can be modelled as:

$$M_{ij} = 1.1\frac{\mu_0}{2}\sqrt{l_i l_j}\left[\left(\frac{2}{\kappa}-\kappa\right)K(\kappa)-\frac{2}{\kappa}E(\kappa)\right], \qquad (10)$$

where $\kappa = 2\sqrt{l_i l_j}/(l_i + l_j)$; and $K(\kappa)$, $E(\kappa)$ represent the complete elliptical integrals of the first and second kind[8]. Consequently the total inductance of the coil antenna is

$$L_{total} = L_{self} + \sum_{i=1}^{N}\sum_{j=1}^{N} M_{i,j} \qquad \in i \neq j \qquad (11)$$

Assuming the antenna conductor material is copper with resistivity ρ, its resistance at ω can be approximated as

$$R = \frac{R_{dc}t}{\delta\left(1-e^{-t/\delta}\right)}, \qquad \delta = \sqrt{2\rho/\mu_0\omega}, \qquad R_{dc} = \ell\rho/tw \qquad (11)$$

t is the thickness of the conductor trace, while ℓ is the total length of the conductor trace, given by

$$\ell = 4Nl_{out} - 4Nw - (w+s)(2N+1)^2 \qquad (12)$$

Thus, antenna inductance and resistance depend on the individual lengths of turns, and the total length of the conductor trace. These lengths can be altered by varying the spacing s between turns, if turn widths w are constant. By implication, the coupling coefficient and Q-factors, defined in equations (5) and (6) respectively, are functions of s. The analysis also suggests the possibility of achieving stronger H-fields by increasing the number of turns, while realizing desired coupling levels for symmetric links with reduced Q-factors by manipulating s. Hence, the design problem becomes

$$\begin{aligned} &\textit{minimize}: && f(s) = \left|k_{ref} - k(s)\right| \\ &\textit{subject to}: && l_{out} \leq l_{max} \\ & && s \geq s_{min} \\ &\textit{with}: && N = N_{ref} \\ & && l_{in} = l_{min} \end{aligned} \qquad (13)$$

k_{ref} is the coupling coefficient that would be obtained by coupling a pair of identical reference coil antennas. The reference coil antenna is a conventional planar square coil antenna with the same number of turns N_{ref} as the desired antenna, but with the turns concentrated along the antenna periphery, and a turn-spacing $s_{\min}$. $l_{\min}$ and $s_{\min}$ determined by installation requirements, and fabrication constraints respectively.

Some assumptions are made with respect to the intended coil antenna design. The installation and fabrication impositions are: $l_{\max} = 35$ mm, $s_{\min} = 0.5$ mm and $l_{\min} = 4$ mm. The antenna is to be implemented on an FR4 printed circuit board, with w = 1 mm, and t = 0.035 mm. The input current level is arbitrary, and the operating frequency is 13.56 MHz. Finally, the symmetric bi-directional link, comprises axially-aligned identical coil antennas placed 30 mm apart, and simultaneously conjugate matched to 50 Ω source and load impedances. Using an S-parameter characterization of the link, the transmission efficiency is computed as[9]

$$\eta(\%) = \left|s_{21}\right|^2 \times 100 \qquad (14)$$

The layout of the modified antenna is realized by a proposed algorithmic solution to equation (13), shown in Table I.

Table 1 Algorithm to distribute turns of planar square coil antenna

```
Start
    get(I, s_min, h, N_ref, ω, δ, μ_0, z, k_ref, l_max, l_min,)
    l_1 = l_min + 2w
    for i = 1 to N_ref
        i = i + 1
        l_i = l_{i-1} + 2s_min + 2w
    end
    compute k(s)_new
    for i = N to 1
        while k(s) < k_ref & l_out < l_max
            l_i = l_i + h
            compute Q, H and k(s)
        end
    end
    T ⇐ l_i,   ∀ i ∈ [1, N]
    return T
Stop
```

The algorithm initializes with the length of the first turn $l_1 = l_{min} + 2w$. Lengths of subsequent $N-1$ turns are initially incrementally laid out from previous turns with spacing s_{min}. The coupling coefficient $k(s)$ between this initial coil antenna and an assumed aligned identical antenna at z is calculated. Subsequent iterations minimize the objective function $f(s)$ by increasing the turn-lengths $l_2, ..., l_N$ by a step-size n = 0.1 mm, without violating the constraint $l_{out} \leq l_{max}$. When $f(s)$ converges at an acceptable minimum, $l_1, ..., l_N$ describe the geometric layout of the proposed coil antenna.

Table 2 Turn-lengths for modelled coil antennas

Design	N	l_1 (mm)	l_2 (mm)	l_3 (mm)	l_4 (mm)	l_5 (mm)
Single-Turn						
	1	35	-	-	-	-
Peripheral-Turns						
	3	29	32	35	-	-
	4	26	29	32	35	-
	5	23	26	29	32	35
Distributed-Turns						
	3	6	31.2	34.2	-	-
	4	6	28.5	31.5	34.5	-
	5	6	25.6	28.6	31.6	34.6

3.0 RESULTS AND DISCUSSION

The algorithm in Table 1 was implemented in MATLAB to derive turn-lengths for 3-, 4- and 5-turn instances. These derived dimensions are listed in Table 2, and illustrated in Figure 7 for a distributed-turn design. The turn layouts listed in Table 2 were then modelled in CST Microwave Studio for full-wave electromagnetic (EM) simulations of axial H-field magnitudes (Figures 1 – 3), Q-factors (Figure 4), reflection coefficients ($|s_{11}|$) (Figure 5), and transmission efficiencies (Figure 6), using 50 Ω port excitations. Transmission efficiency performance was obtained by S-parameter simulations of symmetric bi-directional inductive links.

As is evident in Figures 1 and 2, with 3 and 4 turns, the peripheral-turns design excites stronger axial H-fields than the distributed-turns design. Figure 3, however, shows that with 5 turns, the distributed turn design excites a stronger axial H-field than the conventional peripheral-turns and single-turn designs, and therefore has a potential for greater operational range when used as an interrogator.

Figure 4, however, shows a trade-off with respect to the Q-factor of the distributed 5-turn design. Although there is a reduction in Q-factor of the distributed 5-turns design compared to the peripheral 5-turns design, the H-field enhancement in the distributed turn design has been achieved with a higher Q-factor than the single-turn coil antenna. This higher Q-factor leads to a constriction in the impedance bandwidth of the distributed turns antenna when compared to the single turn design, as shown in Figure 5. The peripheral turns design possesses the smallest impedance bandwidth potential, due to its high Q-factor level. This Q-factor margin, however, becomes beneficial to mitigating the performance degradation due to external loading effects when the same antenna is employed in a symmetric bi-directional inductive coupling link, as shown in Figure 6.

A comparison of Figures 5 and 6 reveals the need for a compromise in the unloaded Q-factor levels of the coil antennas if they are to perform acceptably in both HF-RFID links and bi-directional inductive coupling links. While reduced Q-factor is beneficial to enhanced bandwidth characteristics for HF-RFID coil antennas, as shown in Figure 5, Figure 6 shows that this leads to poor transmission performance in bi-directional links. The single turn coil achieves a transmission efficiency of 58%, as compared to 79% achieved by the 5-turn peripheral turns design. A compromise between these two conflicting requirements has been achieved by the coupling-based approach to distributing the turns of coil antenna, which has achieved a transmission efficiency of 74%.

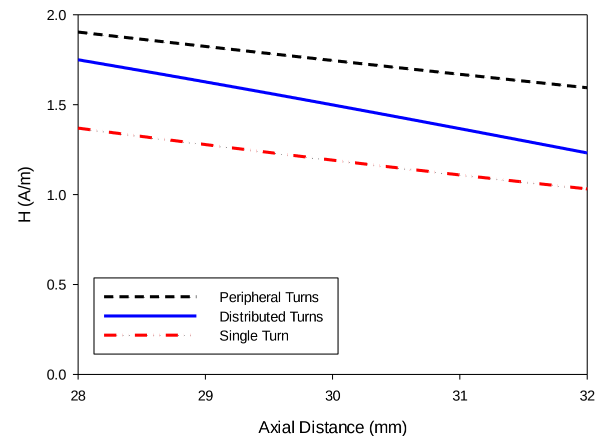


Figure 1 Comparison of H-field magnitude for single-turn and 3-turn peripheral and distributed designs

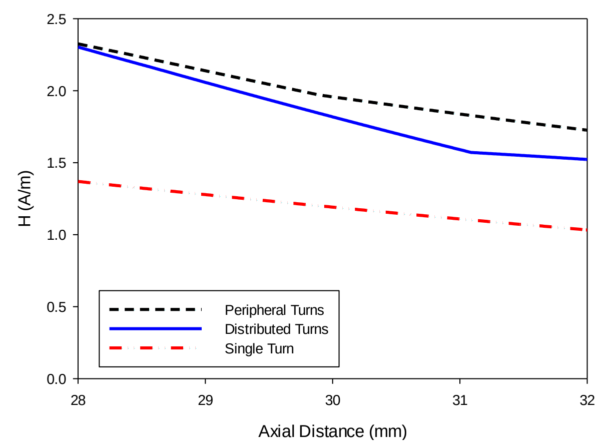


Figure 2 Comparison of H-field magnitude for single-turn and 4-turn peripheral and distributed designs

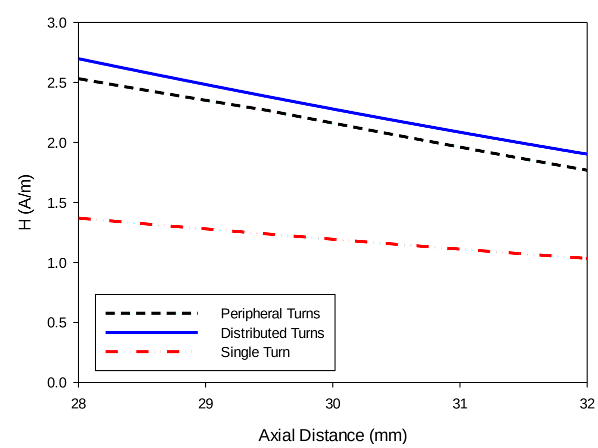


Figure 3 Comparison of H-field magnitude for single-turn and 5-turn peripheral and distributed designs.

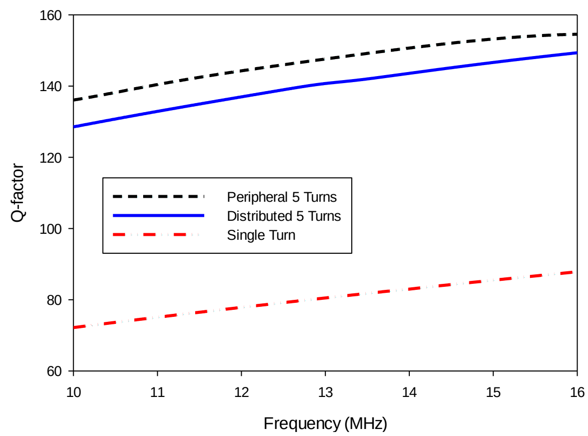


Figure 4 Q-factor comparison between single-turn and 5-turn peripheral and distributed designs.

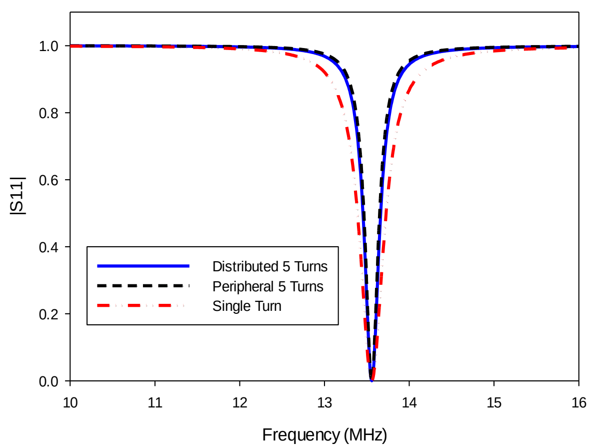


Figure 5 $|S_{11}|$ comparison between single-turn and 5-turn peripheral and distributed designs.

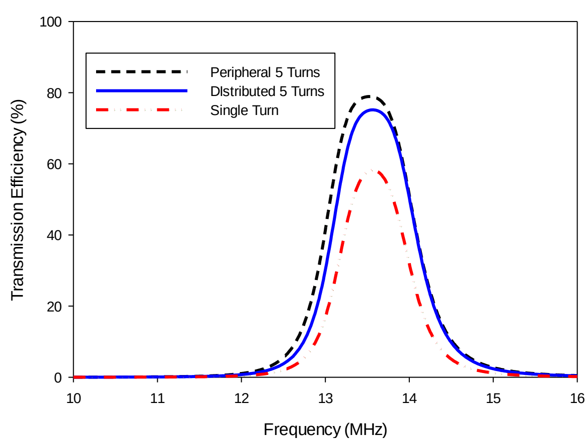


Figure 6 Transmission efficiency comparison between single-turn and 5-turn peripheral and distributed designs.

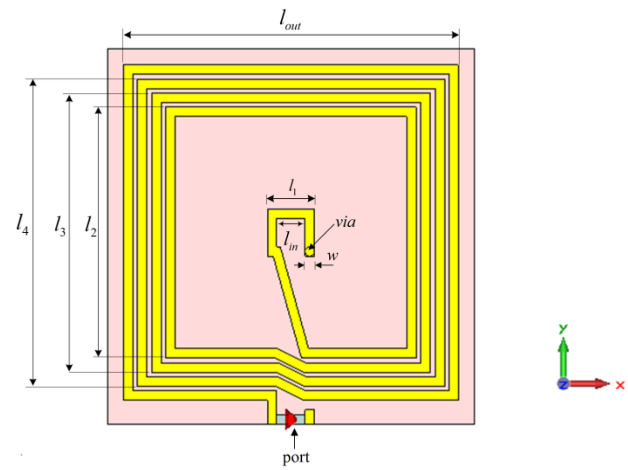


Figure 7 Dimensions of 5-turn distributed-turns coil antenna. For the peripheral-turns design, all turns are concentrated along the coil perimeter and spaced by s_{min}. The single-turn design consists of the outermost turn only.

4.0 CONCLUSION

This paper has presented a numerical study of a coupling-based approach to distributing the turns of planar square coil antennas to enable a dual utility as HF-RFID interrogators and as symmetric bi-directional inductive link terminals. The coupling-based enhancement is used to define the turn-layout of a multi-turn coil antenna to enable the strengthening of the axial H-field for increased HF-RFID interrogator range at less cost of increased Q-factor. As shown by numerical simulations, the achieved Q-factor margin enables enhanced transmission efficiency relative to a minimal Q-factor design when the coil antenna is employed in a symmetric bi-directional inductive coupling link.

Acknowledgement

This work was supported in part by the Collaborative Research in Engineering, Science and Technology (CREST) fund, Malaysia, under grant number: 4B151.

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Appendix

Assuming a current-carrying square filament is centered at the origin of the coordinate system, as shown in Figure A1. By the Biot-Savart law, the magnetic field at **r** due to the current *I* flowing in the square loop is:

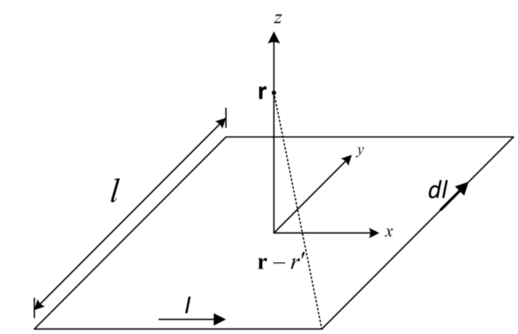

Figure 1 Current-carrying square filament

$$H(\mathbf{r}) = \frac{I}{4\pi}\oint_L \frac{dl \times (\mathbf{r} - r')}{|\mathbf{r} - r'|^3} \tag{A1}$$

Treating the integral in (A1) as a sum of vector contributions from the four currents segments in the square loop leads to

$$H(\mathbf{r}) = \frac{I}{4\pi}\left[\begin{array}{l} \displaystyle\int_{-l/2}^{l/2} \frac{dx'\hat{x} \times \left(-x'\hat{x} + \frac{l}{2}\hat{y} + z\hat{z}\right)}{\left|-x'\hat{x} + \frac{l}{2}\hat{y} + z\hat{z}\right|^3} + \int_{-l/2}^{l/2} \frac{dy'\hat{y} \times \left(\frac{-l}{2}\hat{x} - y'\hat{y} + z\hat{z}\right)}{\left|\frac{-l}{2}\hat{x} - y'\hat{y} + z\hat{z}\right|^3} \\ \displaystyle + \int_{l/2}^{-l/2} \frac{dx'\hat{x} \times \left(-x'\hat{x} - \frac{l}{2}\hat{y} + z\hat{z}\right)}{\left|-x'\hat{x} - \frac{l}{2}\hat{y} + z\hat{z}\right|^3} + \int_{l/2}^{-l/2} \frac{dy'\hat{y} \times \left(\frac{l}{2}\hat{x} - y'\hat{y} + z\hat{z}\right)}{\left|\left(\frac{l}{2}\hat{x} - y'\hat{y} + z\hat{z}\right)\right|^3} \end{array}\right] \tag{A2}$$

However, since the 4 current segments are equal in length, the x- and y-components of their field contributions cancel out, leaving only an aggregate z-component, calculated as:

$$H(z) = 4.\frac{I}{4\pi}\int_{-l/2}^{l/2} \frac{dx'\hat{x} \times \left(-x'\hat{x} + \frac{l}{2}\hat{y} + z\hat{z}\right)}{\left|-x'\hat{x} + \frac{l}{2}\hat{y} + z\hat{z}\right|^3} \cdot \hat{z} \tag{A3}$$

Resolving the vector components leads to a simplification of (A3) as:

$$H(z) = \frac{Il}{2\pi}\int_{-l/2}^{l/2} \frac{dx'}{\sqrt{\left(x'^2 + \frac{l^2}{4} + z^2\right)^3}}, \tag{A4}$$

which becomes:

$$H = \frac{I}{2\pi\left(\frac{z^2}{l^2} + \frac{1}{4}\right)\sqrt{z^2 + \frac{l^2}{2}}}. \tag{A5}$$

If the square filament is modified to consist of *N* very closely-spaced concentric filament turns, then the excited magnetic field along the z-axis can be approximated as:

$$H = \frac{NI}{2\pi\left(\frac{z^2}{l^2} + \frac{1}{4}\right)\sqrt{z^2 + \frac{l^2}{2}}}. \tag{A6}$$