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A MAGNETO-INDUCTIVE LINK BUDGET FOR WIRELESS POWER TRANSFER AND INDUCTIVE COMMUNICATION SYSTEMS

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Abstract—This paper presents a propagation model and inductive link budget based on link equations for chains of inductive loops as the basis for determining the link budget of inductive communication and wireless power transfer systems. The link between the transmitter and receiver is modeled in similar format as in radio frequency systems. The transmitter antenna gain, path loss model and receiver antenna gain are also modeled for the inductive case. This allows the magnetic path loss to be estimated accurately. Also the induced receiver current due to a transmitter voltage can be computed apriori enabling efficient design of inductive links and transceivers.

1. INTRODUCTION

A great deal of propagation models and link budget expressions exist for radio wave propagation at different frequencies and in different terrains. However, when it comes to inductive communication, no propagation models and link budgets similar to Lee [1], Hata, COST 231 and Stanford University Interim (SUI) models [2] exist in current literature. This could be mainly due to the limited research on inductive communications to date. However, interest in magnetic induction (MI) systems operating as wireless power transfer and communication transceivers has increased greatly within the last few years because of their inherent properties. Inductive systems do not interfere with existing traditional electromagnetic wave radiators in most bands. To their advantage magnetic induction communication systems are also not generally affected by the environment. In fact the only parameter in the MI power equation and link budget that

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has to do with the environment is the permeability of the materials in the link and source (sink) which acts to enhance the received signal. Hence the permeability of the medium can be used to enhance the power transfer in the link. Therefore issues such as fading, multipath propagation, interference and noise which plague electromagnetic (EM) systems are not problems in MI communication systems. Inductive systems are however very short range technologies due to rapid decay with distance of the flux created by the varying transmitter current. Applications of inductive methods have become more and more widespread including transcutaneous systems [3], near-field voice communications [4], wireless power transfer [5], data transmission systems [6], underground communications [7, 8], links and communication channels inside integrated circuits [6]. Many of these applications use several coils to either extend the range of the application or to deliver power more efficiently. The more coils used the more the number of equations that must be solved to determine the transfer function of the system. One of such applications where many coils are used is in magneto-inductive waveguides. Magneto-inductive waveguides have recently emerged as a method of extending the range of MI communications systems. The pioneering works of Syms et al. [9–14] and Kalinin et al. [15] have established some of the theories for the MI waveguides. Recently the authors also demonstrated relaying in MI systems [21, 22]. These systems involve arrays of coils arranged as resonant chain networks or in multiple paths to create multipath relay nodes. The solutions to their lumped circuit models normally involve solving systems of simultaneous equations which are prone to mistakes because of the number of variables and equations involved. A system of N resonating nodes requires $(N + 1)$ simultaneous equations to be solved and becomes very difficult when N is large. The objective of this paper is to propose a fast solutio method in other to simplify this rigour.

The methods of link budgets for traditional radio frequency (RF) communications are well established and understood. A link budget presents a summary of how the transmitted power is spent in the communication chain between the transmitter and receiver. System gains are presented as positive values and losses as negative values in decibels. Over the last couple of years attempts have been made towards developing link budget expressions for inductive systems as given in [15–22]. Lack of consistency in the formulations of the link budget expressions which limits their use motivates this paper.

Inductive links to date have been developed as chain networks in which one loop induces flux on its neighbor until the data flux is received by the last loop to which the load is connected. Usually the systems have line-of-site of each other and are aligned along the common axis of the coils. Such systems may therefore be modeled as resonating magneto-inductive waveguides. Therefore in this paper the analysis leading to the elegant expression for the inductive link budget has used this approach as depicted in Figure 1. The rest of the paper is organized as follows. Section 2 develops expressions for inductive links using the magneto-inductive waveguide formulation. Simulations of the links are presented in the same section. In Section 3, the magnetic link equation is derived and used to propose a link budget expression for low coupling applications. In Section 4, methods for assessing the efficiency of the magnetic link are proposed with conclusions drawn in Section 5.

2. INDUCTIVE LINKS

In its simplest form, a one section peer-to-peer magnetic field coupling system is used for wireless power transfer and communication in the traditional MI system with no options of range extension (Figure 1). This model has been well analyzed and discussed by Agbinya et al. [6]. The lumped circuit model of the system is also shown in Figure 1. An N-section waveguide extends this simple formulation by using N resonating coils or split-ring resonators. Figure 1 shows the multiple coil array version of the system. We assume each coil is loaded with a capacitor to resonate and the receiver has a load Z_L. The current in each loop n is therefore I_n. We assume only nearest neighbor interaction in which only currents in node $n-1$ and $n+1$ affect node n. we also assume that the N nodes are resonating at frequency ω_0. Node n has impedance Z_n and current I_n. Node $n=1$ is excited with input voltage and the rest couple magnetic fields from one to the other until the receiver node is reached. The intermediate nodes are passive but the receiver has load impedance Z_L.

2.1. Power Relations in Inductive Links

The purpose of this section is to establish a system equation which is repeatable and easily usable in a form similar to the propagation equation in basic electromagnetic communications systems. To do this, let us consider a peer-to-peer inductive communication consisting of two loops, a transmitter and receiver loops. The governing equations for the system (Figure 1) are:

$$Z_1 I_1 + j\omega M_{12} I_2 = U_s \tag{1}$$

$$Z_2 I_2 + j\omega M_{12} I_1 = 0 \qquad M_{ij} = k_{ij}\sqrt{L_i L_j} \qquad 0 \le k_{ij} \le 1 \tag{2}$$

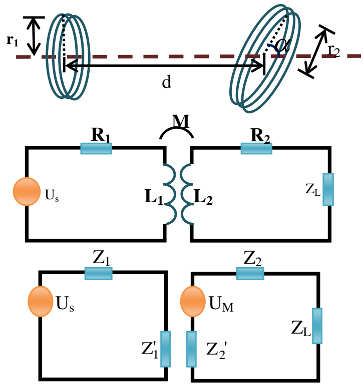

Figure 1. MI communication system.

where the transmitter and receiver loop impedances are Z_1 and Z_2 with currents flowing through them as I_1 and I_2 respectively. The mutual inductance and the coupling coefficient between them written in general terms are M_{ij} and k_{ij} respectively ($i = 1$, $j = 2$). Let the transmitter parameters have index '1' and index '2' be reserved for the receiver variables. The impedances Z_2' refer to the influences of the transmitter on the receiver and of the receiver on the transmitter Z_1'.

The voltage developed across the receiver load due to inductive action is given by the expression

$$V_L = I_2 Z_L \tag{3}$$

where in general for a system consisting of multiple intermediate (or relay) nodes each loop impedance is

$$Z_n = R_n + j\omega L_n + \frac{1}{j\omega C_n} \tag{4}$$

The impedance of the receiver when the load impedance is considered is

$$Z_r = R_2 + j\omega L_2 + \frac{1}{j\omega C_2} + Z_L \tag{5}$$

By solving Equations (1) to (5) simultaneously, the ratio of the load voltage to that of the source is given by the expression:

$$G_v = \frac{V_L}{U_S} = \frac{-j\omega M_{12} Z_L}{\omega^2 M_{12}^2 + Z_1 Z_2} \tag{6}$$

The gain function G_v relates the input voltage (U_S) to the voltage V_L developed across the receiver load Z_L. It relates the system parameters for the transmitter and receiver including the link between them. This expression may also be written in terms of the current induced in the receiver coil as:

$$G_v(\omega) = \frac{I_L}{U_S} = \frac{-j\omega M_{12}}{\omega^2 M_{12}^2 + Z_1 Z_2} \tag{7}$$

where $I_L = I_2 = V_L/Z_L = G_v U_S$. The power delivered to the receiver load is the product $I_L^2 Z_L = I_2 I_2^* Z_L$.

2.1.1. Approximate Wireless Power Transfer Equation for a Link

Equation (7) is a simple case because it involves solving only three equations but once the number of loops increase and for N loops $(N+1)$ equations are required, solving the equations to obtain $G_v(\omega)$ becomes a lot more difficult and time consuming. A preferred approach is to develop a link equation which relates multiple loops in terms of the input transmitter voltage to the inductive current flowing through the load impedance. Let the quality factors of the transmitter and receiver coils be $Q_1 = \omega L_1/R_1$; $Q_2 = \omega L_2/R_2$ and replacing the mutual inductance with an equivalent expression containing the quality factors, the expression for the inductive system gain at resonance becomes

$$\frac{I_L}{U_S} = \frac{-jk_{12}\sqrt{Q_1 Q_2}}{\sqrt{R_1 R_2}\left(k_{12}^2 Q_1 Q_2 + 1\right)} \tag{8}$$

At low coupling and small quality factors (Q), the inequalities $k_{12}^2 Q_1 Q_2 \ll 1$ or $k_{12}^2 \ll 1/Q_1 Q_2$ hold. Therefore the inductive transfer function at resonance reduces to.

$$\left|\frac{I_L}{U_S}\right| = \frac{k_{12}\sqrt{Q_1 Q_2}}{\sqrt{R_1 R_2}} = \frac{k_{ij}\sqrt{Q_i Q_j}}{\sqrt{R_i R_j}} \tag{9}$$

We have simulated Equations (8) and (9) in Matlab. Figures 2(a) and (b) show the simulation for the low coupling approximation when the two coils have identical $Q = 2$ and $Q = 10$ respectively and $k = 0.1$. The approximation is very accurate for low k and low Q but starts to deviate for higher Q. Figure 2(c) shows the variation of the transfer function with Q and plotted as a function of the logarithm of the magnitude of the numerator of Equation (8). Using the horizontal axis from these figures, it is apparent that the approximation is valid at high Q when the numerator is nearly equal to the denominator. Very high Q causes a deviation from the general solution.

2.1.2. Approximate Wireless Power Transfer Link Equations

The use of multiple inductive loops (Figure 4) in a chain network is more popular than the simple two loops system. Increasing the

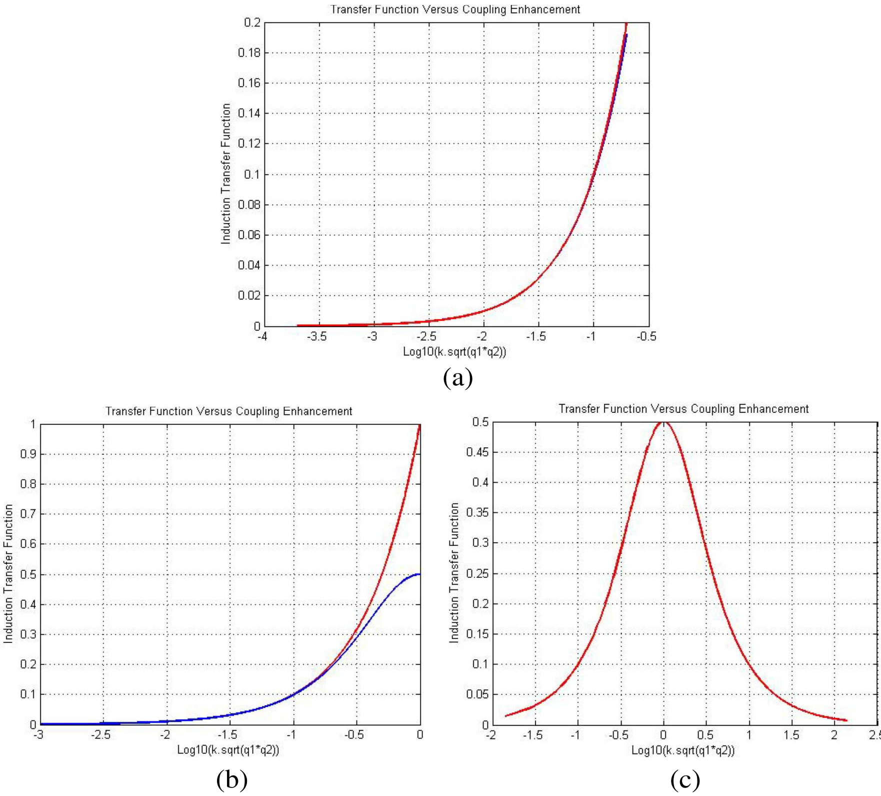

Figure 2. Compared transfer functions with/without approximation ($N = 2$). (a) Compared transfer function at low Q (red curve is with approximation; blue curve has no approximation. (b) Compared transfer function at high Q, with approximation (read curve); without approximation (blue curve). (c) Variation of transfer function with Q.

number of loops is used to improve the power transfer efficiency and for extending the communication range. Consider a three node system with the following Kirchhoff voltage law (KVL) equations:

$$\begin{aligned} &Z_1 I_1 + j\omega M_{12} I_2 = U_s \\ &Z_2 I_2 + j\omega M_{12} I_1 + j\omega M_{23} I_3 = 0 \\ &Z_3 I_3 + j\omega M_{23} I_2 = 0 \\ &V_L = I_3 Z_L \end{aligned} \tag{10}$$

Solving the simultaneous Equation (10) results to the expression:

$$\frac{V_L}{U_S} = \frac{\omega^2 M_{12} M_{23} Z_L}{\omega^2 \left(M_{12}^2 Z_3 + M_{23}^2 Z_1 \right) + Z_1 Z_2 Z_3} \tag{11}$$

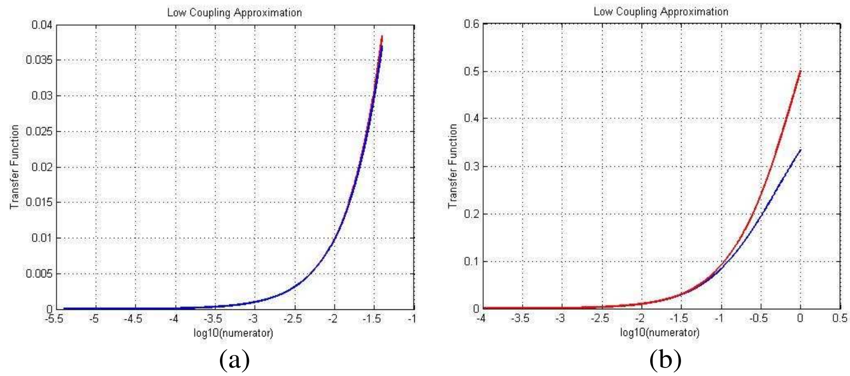

Figure 3. Power transfer functions with approximation ($N = 3$). (a) Power transfer function when $Q = 2$; Red = approximation; Blue = no approximation. (b) Power transfer function when $Q = 10$; Red = approximation; Blue = no approximation.

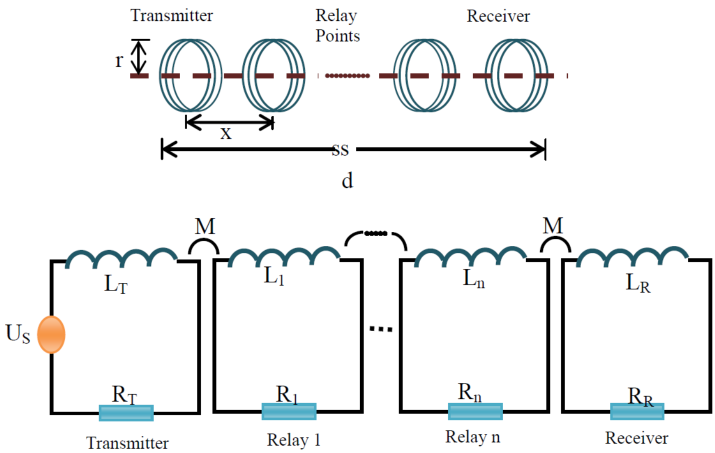

Figure 4. Magnetic waveguide and its circuit model.

At resonance this equation can be simplified and becomes

$$\frac{I_L}{U_S} = \frac{k_{12}k_{23}Q_2\sqrt{Q_1Q_3}}{\sqrt{R_1R_3}\left[1 + k_{12}^2Q_1Q_2 + k_{23}^2Q_2Q_3\right]} \tag{12}$$

As in Equation (8), we consider low coupling when $k_{12}^2Q_1Q_2 + k_{23}^2Q_2Q_3 \ll 1$ or $Q_2 \ll 1/\left(k_{12}^2Q_1 + k_{23}^2Q_3\right)$ and writing $k_{ij} = k_{i,j}$ the low coupling transfer function becomes

$$\left|\frac{I_L}{U_S}\right| = \frac{k_{12}k_{23}Q_2\sqrt{Q_1Q_3}}{\sqrt{R_1R_3}} = \frac{\sqrt{Q_1Q_N}\prod\limits_{i=1}^{N-1}k_{i,i+1}\prod\limits_{i=1}^{N-2}Q_{i+1}}{\sqrt{R_1R_N}} \tag{13}$$

To demonstrate the effectiveness of the approximations, simulations were undertaken using Matlab. Equation (12) the case for no approximation and Equation (13) with approximation were used. The red lines in Figure 3 show the case when no approximations were used. The blue curves are results with approximation. The approximation for $k = 0.1$ and $Q = 2$ is held at $N = 3$ with slight deviations only for higher $Q = 10$. High Q values are normally used in wireless power transfer systems and the higher the Q the more the deviations in the curves shown in Figure 3. These approximations are therefore more suited to low Q systems as in inductive communications. A correction term $(k_{12}^2Q_1Q_2)$ should be used in the denominator of Equation (13) to reduce the deviation at high Q. The correction changes the denominator to $\sqrt{R_1R_3}\left(1 + k_{12}^2Q_1Q_2\right)$.

We extend the system one more time to demonstrate the concept further for $N = 4$. In this case the system of equations is

$$\begin{aligned}
&Z_1I_1 + j\omega M_{12}I_2 = U_s\\
&Z_2I_2 + j\omega M_{12}I_1 + j\omega M_{23}I_3 = 0\\
&Z_3I_3 + j\omega M_{23}I_2 + j\omega M_{34}I_4 = 0\\
&Z_4I_4 + j\omega M_{34}I_3 = 0\\
&V_L = I_4Z_L
\end{aligned} \tag{14}$$

This has the solution

$$\frac{V_L}{U_S} = \frac{j\omega^3M_{12}M_{23}M_{34}Z_L}{Z_1Z_2Z_3Z_4 + \omega^4M_{12}^2M_{34}^2 + \omega^2\begin{pmatrix}M_{12}^2Z_3Z_4 + M_{23}^2Z_1Z_4\\ +M_{34}^2Z_1Z_2\end{pmatrix}} \tag{15}$$

At resonance it reduces to

$$\left|\frac{I_L}{U_S}\right| = \frac{Q_2Q_3k_{12}k_{23}k_{34}\sqrt{Q_1Q_4}}{\sqrt{R_1R_4}\begin{bmatrix}1 + \left(k_{12}^2Q_1Q_+k_{23}^2Q_2Q_3 + k_{34}^2Q_3Q_4\right)\\ +k_{12}^2k_{34}^2Q_1Q_2Q_3Q_4\end{bmatrix}} \tag{16}$$

The low coupling approximation when $N = 4$ with $t(k_{12}^2Q_1Q_2 + k_{23}^2Q_2Q_3 + k_{34}^2Q_3Q_4) + k_{12}^2k_{34}^2Q_1Q_2Q_3Q_4 \ll 1$ is

$$\left|\frac{I_L}{U_S}\right| = \frac{k_{12}k_{23}k_{34}Q_2Q_3\sqrt{Q_1Q_4}}{\sqrt{R_1R_4}} = \frac{\sqrt{Q_1Q_4}\prod\limits_{i=1}^{3}k_{i,i+1}\prod\limits_{i=1}^{2}Q_{i+1}}{\sqrt{R_1R_4}} \tag{17}$$

Normally even when the loops are arranged equidistant from each other, the strongest coupling is between the first and second loop with coefficient k_{12} and the rest of the coefficients become smaller and smaller towards the receiver.

To accommodate for high coupling situations as in wireless power transfer, correction terms may also be progressively added to the denominator of Equation (17).

3. INDUCTIVE LINK EQUATION

The performance of an inductive system is conditioned upon reducing losses in the transmitter, the receiver and the channel. Both the receiver and transmitter suffer from resistive losses. The previous section has provided the basis for assessing the performance of the inductive link. Losses in the transmitter and receiver will be discussed at the end of this section. From Equations (9), (13) and (17) a general link equation for N loops in a chain can be derived using the low coupling approximation and nearest neighbor interaction. Using Equations (9), (13) and (17) we can show that the link equation and hence the link efficiency η_{TR} is given by the following expression

$$\eta_{TR} = \left|\frac{I_L}{U_S}\right| = \frac{\sqrt{Q_1 Q_N}\prod_{i=1}^{N-1} k_{i,i+1}\prod_{i=1}^{N-2} Q_{i+1}}{\sqrt{R_1 R_N}} \tag{18}$$

This is a general expression which holds for all N, provided the approximations are made in the denominators for the solutions to the simultaneous KVL equations. Equation (18) represents a general simplification of the link equation when multiple loops are involved. To use the equation, it is required that the number of loops N be selected including the electrical dimensions of each loop. The electrical dimensions include the radius r_i of each loop, number of turns N, the resistance of each loop R_i, the distance between the loops l which are then used to compute the quality factors of the loops. Then select the excitation voltage U_S and the load impedance. The load could be a sensor or a device being driven by the array of inductive loops. Therefore the link budget equation for the induced load voltage $V_L = I_L Z_L$ is proportional to the square of the voltage across the load impedance and is given in decibels for N loops as:

$$V_L\,(\text{dB}) = 20\log_{10} U_S + 10\log_{10}(Q_1 Q_N) + 20\sum_{i=1}^{N-1}\log_{10}(k_{i,i+1})$$

$$+20\sum_{i=1}^{N-2}\log_{10}\left(Q_{i+1}\right)-10\log_{10}\left(R_1R_N\right) \qquad (19a)$$

Generally $k_{i,i+1} \ll 1$. Therefore the third term in Equation (19a) is negative and the equation is written in decibels as

$$\begin{aligned} V_L\,(\text{dB}) \;=\; & U_S\,(\text{dB})+Q_{1N}\,(\text{dB})-k_{i,i+1}\,(\text{dB}) \\ & +Q_{i+1}\,(\text{dB})-R_{1N}\,(\text{dB}) \end{aligned} \qquad (19b)$$

We have used a minus sign following the tradition in RF systems for using negative sign for the path loss model. This is because the individual coupling coefficients are less that one. Note that the index variations for the coupling coefficients $k_{i,i+1}$, $1 \leq i \leq N-1$ and for the quality factors Q, $1 \leq i \leq N-2$.

What do Equations (19a) and (19b) represent in terms of wireless power transfer and communication using inductive nodes? By defining Q as the gain of a node, the general equation has the following interpretation as in traditional RF systems. The product $\sqrt{Q_1Q_N}$ is the gain of the transmitter and receiver stages. This is similar to the product of the transmitter and receiver antenna gains in RF systems. The quantity $\prod_{i=1}^{N-2} Q_{i+1}$ is the product of the gains of the intermediate or relay stages. This is also similar to the product of the gains of the intermediate stages in a transponder system. The term $\prod_{i=1}^{N-1} k_{i,i+1}$ is the path loss of the general magnetic channel. Thus we model the overall system of N resonating inductive loops as in Figure 5.

The resistors R_1 and R_N are the inherent Ohmic losses of the transmitter and receiver stages. The inductive system has a channel with path loss (gain) equal to $G_c = \prod_{i=1}^{N-1} k_{i,i+1} \prod_{i=1}^{N-2} Q_{i+1}$. If as in transcutaneous systems applications or in embedded biomedical systems when biological tissues are part and parcel of the channel between the transmitter and receiver, and if the biological channel has impedance Z_b, the channel gain is modified to $G_c = Z_b \prod_{i=1}^{N-1} k_{i,i+1} \prod_{i=1}^{N-2} Q_{i+1}$ and the channel becomes a function of frequency and leads to further coupling losses between the transmitter and receiver. Most importantly, Equations (19a) and (19b) may be used to estimate the various system parameters during the design process to enable a required inductive current in the receiver load of the Nth loop.

Consider as an example an inductive waveguide link system

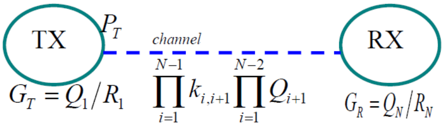

Figure 5. Peer-to-peer inductive communication system.

Table 1. Inductive link parameters.

Parameter	Value	dB
U_S	10 volts	20
$Q_1 = Q_N$	10	20
Q_{i+1}	10	40
$R_1 = R_N$	$1.6\,\Omega$	-4.08
N	4	

parameters of four sections ($N = 4$) as in Table 1. What would the receiver voltage be with various values of the coupling coefficients?

From Equation (19b), the link budget equation reduces to

$$V_L\,(\text{dB}) = 75.92 - 20\sum_{i=1}^{N-1}\log_{10}(k_{i,i+1}) = 75.92 - k_{i,i+1}\,(\text{dB}) \qquad (20)$$

$k_{i,i+1}\,(\text{dB}) = 20\sum_{i=1}^{N-1}\log_{10}(k_{i,i+1})$. This link budget equation places emphasis on the coupling coefficients of the individual sections of the link. From Equation (20) when $N = 4$, considering low coupling case with sections of equal length and identical coils resonating at the same frequency, the load voltage in decibels is found. Let $k_{i,i+1} = 10^{-3}$; $1 \le i \le 3$ the receiver voltage is (-104.08) dB or $6.25\,\mu$V. When $k_{i,i+1} = 10^{-1}$; $1 \le i \le 3$, the receiver voltage is approximately 6.25 volts. Therefore the link Equations (19a) to (20) provide easy method of computing the required system variables. The constant 75.92 in Equation (20) will be slightly higher for large Q systems.

3.1. Link Budget with TX and RX Losses

In general the power transfer efficiency of an inductive system contains three distinct factors. These are the losses in the transmitter antenna and the losses in the receiver antenna. Hence the link equations also should include these losses in the transmitter and receiver circuits. Given a transmitter source resistance R_S, transmitter inductor wire resistance R_1, receiver inductor resistance R_N and load resistance R_L, the transmitter efficiency is η_T and the receiver efficiency is η_R. Hence the the overall system link efficiency is,

$$\eta = \eta_T \eta_{TR} \eta_R; \quad \begin{cases} \eta_T = \frac{R_S}{R_S + R_1} \\ \eta_R = \frac{R_L}{R_L + R_N} \end{cases} \tag{21}$$

With this correction, the link Equations (19a) to (20) need to be modified to include these terms. Since $\eta_T < 1$ and $\eta_R < 1$, the overall system link budget expression is obtained by modifying Equation (19b) to be:

$$\begin{aligned} V_L\,(\text{dB}) \;=\; & -\eta_T + U_S\,(\text{dB}) + Q_{1N}\,(\text{dB}) - k_{i,i+1}\,(\text{dB}) + Q_{i+1}\,(\text{dB}) \\ & -R_{1N}\,(\text{dB}) - \eta_R \end{aligned} \tag{22}$$

The two new terms are marginal corrections which should be implemented to account for the resistive losses in the transmitter and receiver coils.

3.2. Correction Terms

As in RF propagation models and link budgets, progressive correction terms may be added to the denominator of the transfer function equation and hence to the link budget equation in decibels. These corrections apply for high Q when only the first loop is excited with input voltage at frequency ω. For example, when $N = 2$ the correction term is $k_{12}^2 Q_1 Q_2$. When $N = 3$, there are two terms $k_{12}^2 Q_1 Q_2 + k_{23}^2 Q_2 Q_3$ that could be used for corrections and are functions of k_{12}^2 and k_{23}^2. If the nodes are distributed evenly in a chain network so that the distance between nodes is constant x, then $k_{23}^2 < k_{12}^2$ and to improve upon the approximation the term $k_{12}^2 Q_1 Q_2$ may be added as correction. For $N > 3$, progressive corrections may be made for higher $k(Q)$ based on adding terms in k_{ij}^2 where i, and j are integers.

4. CONCLUSIONS

We have developed an accurate expression for link budgets of inductive systems. The algorithm presented enables accurate estimation of inductive loop system gains. They could be used for designing and assessing the performances of embedded biomedical data and wireless power transfer, personal area networks and communications underground. The link budget applies for all N and eliminates the need to solve a large system of equations, provided the coupling coefficients and the quality factors of the loops are computed and known. Progressive corrections to account for larger k and Q can be made for large k and Q applications.

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