

A Novel Simultaneous Wireless Power and Information Transfer Strategy Based on Wave Trappers and Partial Power Coils

Qingqing Yuan

yuanqq@usst.edu.cn

Shanghai University of Science and Technology

Pu Wang

Shanghai University of Science and Technology

Xunyu Zhou

Shanghai University of Science and Technology

Kai Xue

Shanghai University of Science and Technology

Dirui Yang

Shanghai University of Science and Technology

Research Article

Keywords: Full duplex communication SWPIT, ASK modulation, Wave trapper, Partial power coil SS Topology

Posted Date: August 13th, 2026

DOI: <https://doi.org/10.21203/rs.3.rs-9977899/v1>

License:   This work is licensed under a Creative Commons Attribution 4.0 International License.
[Read Full License](#)

Additional Declarations: No competing interests reported.

A Novel Simultaneous Wireless Power and Information Transfer Strategy Based on Wave Trappers and Partial Power Coils

Qingqing Yuan^{1*} · Pu Wang¹ · Xunyu Zhou¹ · Kai Xue¹ · Dirui Yang¹

Abstract

Traditional simultaneous wireless power and information transfer (SWPIT) systems face the challenge of balancing power transfer efficiency with high-speed information transmission. Additionally, interference between the power and information transmission paths often reduces system integration and increases design complexity. To address these issues, this paper proposes a novel full-duplex SWPIT strategy that uses LC wave trappers and partial power coils to physically isolate the power and information transmission paths, minimizing mutual interference between power and information transfer. The proposed method employs ASK modulation for information transmission and an SS topology for power transfer, simplifying both the information modulation and power transfer processes while maintaining high performance. Experimental results show that with 54.7 W of power transmission, the system achieves a forward information transmission rate of 200 kbit/s and a backward transmission rate of 600 kbit/s with low bit error rates and efficient full-duplex communication. Compared with existing SWPIT solutions, the proposed method offers improved integration, scalability, and cost-effectiveness, making it a promising solution for wireless power transfer and intelligent terminal communication.

Keywords: Full duplex communication SWPIT · ASK modulation · Wave trapper · Partial power coil SS Topology

1 Introduction

Wireless Power Transfer (WPT) delivers power without physical connections, enhancing safety and flexibility [1,2]. It is widely used in wireless scenarios such as implantable medical devices and electric vehicles. Practical WPT systems demand both high-efficiency power transfer and reliable bidirectional communication for operational data exchange [3].

Traditional wireless communication technologies such as ZigBee and Bluetooth often introduce significant transmission delays and complex pairing procedures when applied to WPT systems [4]. In contrast, simultaneous wireless power and information transfer (SWPIT) integrates power and data transmission into a single link, reducing system complexity while improving efficiency and compactness [5,6].

Existing SWPIT systems fall into dual-path and shared-coupled-path structures. Dual-path systems use multiple coil sets, increasing size and cross-coupling. Shared-coupled-path systems use the same magnetic link for power and data, enabling a more compact design. Two main methods are widely investigated: power modulation and carrier injection. Power modulation modulates the power carrier directly, with simple implementation but low data rates and half-duplex operation [7,8].

Carrier injection superimposes a high-frequency information carrier on the power link, enabling higher data rates via frequency separation. OOK, BFSK and BASK modulations have been widely adopted [9–12]. Although communication performance is improved, most systems remain half-duplex or limited in efficiency, stability and transmission range. Other integrated schemes achieve moderate efficiency and power but are still restricted to half-duplex operation, restricting real-time bidirectional interaction [13].

Recent studies have investigated full-duplex SWPIT systems to enhance communication performance. However, most rely on complex duplex structures, high-frequency carriers or advanced modulation, increasing complexity, cost and design difficulty [14–17]. Balancing simplicity,

* Qingqing Yuan
yuanqq@usst.edu.cn

¹ Department of Electrical Engineering, School of Electrical and Mechanical Engineering, Shanghai University of Science and Technology, Shanghai, China

communication performance and power efficiency remains a key challenge. Thus, developing efficient, low-complexity full-duplex SWPIT systems is of great importance.

To address the limitations of existing full-duplex SWPIT systems in terms of structural complexity, cost, and device dependency, this paper proposes a novel full-duplex SWPIT system based on LC wave trappers and partial power coils. The proposed topology uses LC networks instead of transformers or duplexers, thereby reducing system complexity and implementation cost. In the proposed scheme, power and forward information are transmitted through the main coupled coils, whereas backward information is injected through the inner partial coil, which effectively suppresses carrier interference. Moreover, the LC wave trappers isolate the forward information channel from the power path, while the backward path relies on the band-pass characteristic of the resonant network for high-frequency isolation. Experimental results verify that the proposed system can simultaneously transmit power and bidirectional information at three different frequencies and exhibits strong anti-interference capability. Under 54.7 W power transfer, 200 kbit/s forward transmission, and 600 kbit/s backward transmission, stable full duplex communication is achieved.

2 System Structure and Working Principle

2.1 Coil Structure

To realize physical separation between the forward, backward, and power paths, the proposed topology introduces a partial-coil structure with wave trappers. The overall topology and the roles of the three transmission paths are first introduced, followed by the partial-coil implementation. The partial coil structures used in this paper are shown in Fig. 1.

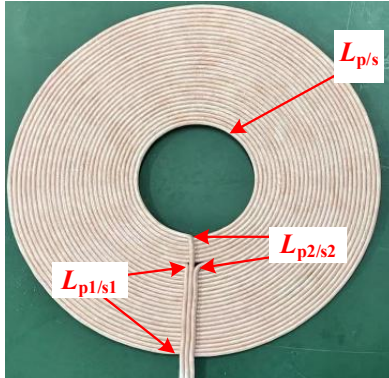


Fig. 1. Partial Coil Structure.

Here, L_{p1} and L_{s1} represent the equivalent inductances of the outer coils on the primary and secondary sides, respectively; and L_{p2} and L_{s2} denote the equivalent inductances of the inner coils on the primary and secondary sides, respectively.

2.2 System Structure

Since the SS compensation topology can readily achieve maximum power transfer with high efficiency, the proposed full duplex SWPIT system is developed on the basis of an SS compensated wireless power transfer topology. The proposed system topology is shown in Fig. 2.

As shown in Fig. 2, this proposed system consists of three components: (1) Power transmission circuit; (2) Downlink communication circuit; (3) Uplink communication circuit. For the power transmission path, the DC power source E_{dc} is firstly converted into a high-frequency AC square-wave voltage by the high-frequency full-bridge inverter (S_1 - S_4). Then this AC voltage is transmitted via the primary-side resonant structure (C_{p1} , $L_p=L_{p1}+L_{p2}$). On the secondary side, the resonant structure (C_{s1} , $L_s=L_{s1}+L_{s2}$) is used to receive the power, which is then subsequently supplied to the load R_L through the rectifier filter circuit (D_1 - D_4 , C_L). Both forward and backward information are communicated through the information branch within the power information transmission circuit, which is the main content introduced in the following text.

In Fig. 2, there are four wave trappers (as shown specifically in Fig. 3) to realize the information communication.

Wave trappers (a) and (b) are connected in series within the power transmission branch, and their resonant frequency is set to match the high frequency information carrier AC₁'s frequency f_i . Consequently, these circuits behave as open circuits at the carrier frequency f_i , functioning as blockers to confine the information carrier's transmission within the coil. Simulation verification confirms that without the wave trappers (a) and (b), the forward information carrier AC₁ would suffer excessive losses and become undetectable at the receiving resistor R_L .

Since the information carrier frequency usually exceeds the power transmission frequency, the wave trappers exhibit inductive impedance at the power transmission frequency. Therefore, capacitors C_{1lc} , C_{12c} must be connected in series to achieve zero total impedance, preventing the wave trappers from interfering with power transmission.

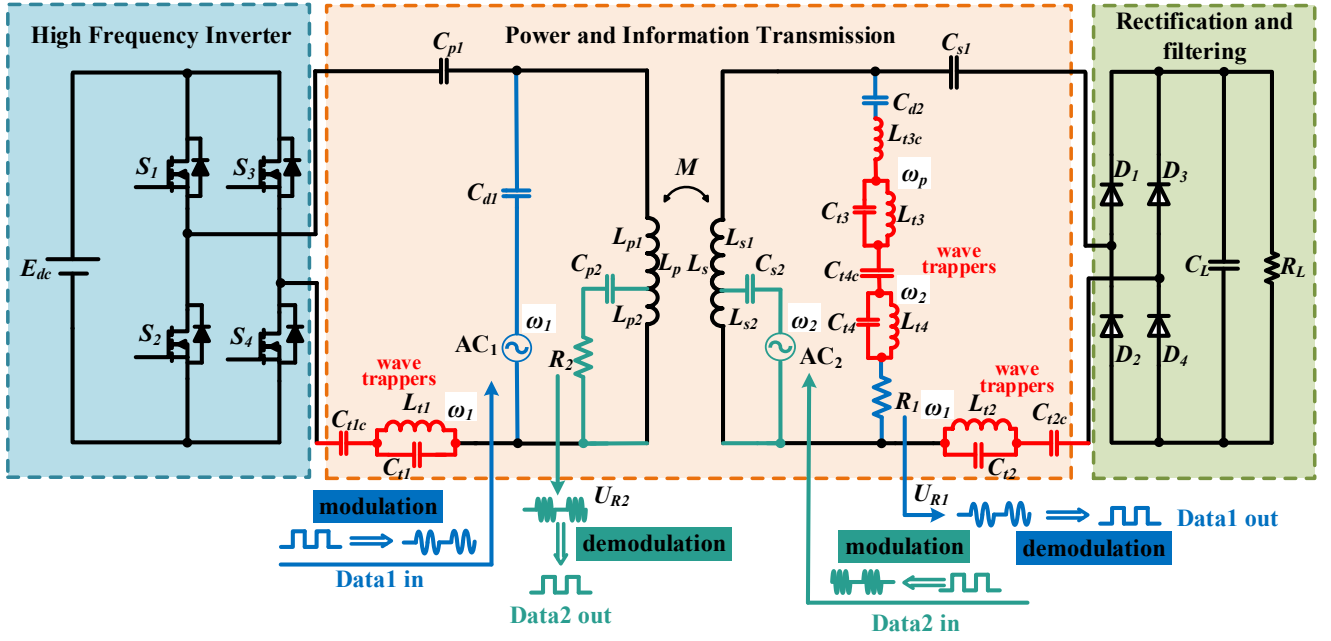


Fig. 2. System structure.

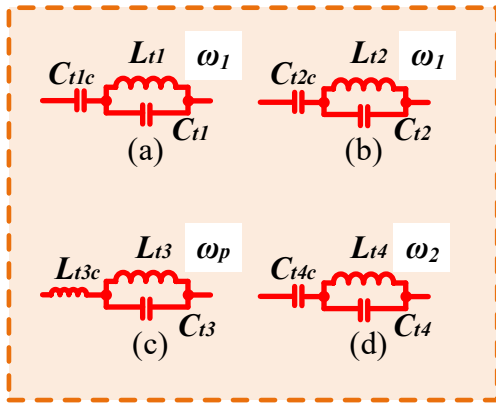


Fig. 3. Wave Trappers.

In Fig. 3(c) and Fig. 3(d), the parallel resonant frequencies of L_{13} , C_{13} , L_{14} , C_{14} correspond to the power transmission frequency f_p and the backward information frequency f_2 , respectively. These two wave trappers are connected in series within the forward information detection circuit to prevent power and backward information carrier signals from entering this branch during transmission, which would cause significant interference with the voltage U_{R1} across the receiving resistor R_I and hinder accurate information demodulation.

Similarly, the series inductor L_{13c} and capacitor C_{14c} ensure the entire circuit has zero impedance at f_1 to prevent interference with information detection. Capacitors C_{11c} , C_{p1} and C_{12c} , C_{s1} should satisfy the series capacitance formula. To simplify circuit analysis, symmetrical parameters can be used in the design, meaning that $L_{11}=L_{12}$, $C_{11}=C_{12}$ and $C_{11c}=C_{12c}$. Then, f_1 , f_2 , and f_p can be represented as

$$\begin{cases} f_1 = \frac{1}{2\pi\sqrt{L_{11}C_{11}}} = \frac{1}{2\pi\sqrt{L_{12}C_{12}}} \\ f_p = \frac{1}{2\pi\sqrt{L_{13}C_{13}}}, \quad f_2 = \frac{1}{2\pi\sqrt{L_{14}C_{14}}} \end{cases} \quad (1)$$

2.3 Power Transmission

The output voltage of a full-bridge inverter controlled by two complementary PWM signals with dead zones is

$$\begin{cases} \omega_p = \frac{1}{\sqrt{L_p C_{p1}}} = \frac{1}{\sqrt{L_s C_{s1}}} \\ u_{inv} \approx \frac{4E_{dc}}{\pi} \sin(\omega_p t) \end{cases} \quad (2)$$

Where ω_p denotes the fundamental angular frequency. Since f_1 and f_2 are much higher than f_p , the information branches behave as open circuits at f_p , whereas the wave trappers (a) and (b) with their compensations act as short circuits.

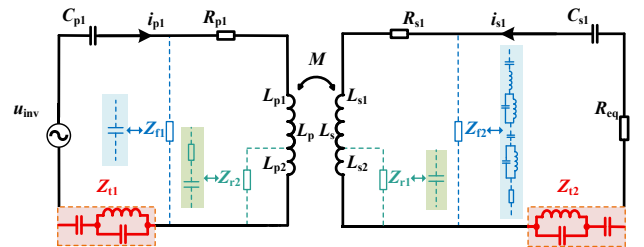


Fig. 4. Equivalent Circuit Diagram of Power Transmission.

Hence, the circuit is equivalent to an SS-compensated WPT system (as shown in Fig. 4), where R_{eq} represents the combined resistance of the rectifier and the load R_L , which can be given as $R_{eq}=(8R_L)/\pi^2$. In Fig. 4, Z_{f1} , Z_{f2} , Z_{r1} , Z_{r2} denote the equivalent impedances of corresponding

information transmission paths; Z_{t1} , Z_{t2} are equivalent impedances of band rejection circuits; R_{p1} , and R_{s1} represent the transmitter-side and receiver side internal resistances of the coil, respectively. Where

$$\begin{cases} Z_{f1} = \frac{1}{j\omega C_{d1}}, Z_{f2} = \frac{1}{j\omega C_{d2}} + j\omega L_{t3c} \\ + \frac{j\omega L_{t3}}{1 - \omega^2 L_{t3} C_{t3}} + \frac{1}{j\omega C_{t4c}} + \frac{j\omega L_{t4}}{1 - \omega^2 L_{t4} C_{t4}} + R_1 \\ Z_{r1} = \frac{1}{j\omega C_{s2}}, Z_{r2} = \frac{1}{j\omega C_{p2}} + R_2 \\ Z_{t1} = \frac{1}{j\omega C_{t1c}} + \frac{j\omega L_{t1}}{1 - \omega^2 L_{t1} C_{t1}} \\ Z_{t2} = \frac{1}{j\omega C_{t2c}} + \frac{j\omega L_{t2}}{1 - \omega^2 L_{t2} C_{t2}} \end{cases} \quad (3)$$

At system resonance, the input impedance and the primary and secondary currents are

$$\begin{cases} Z_{in} = R_{p1} + \frac{(\omega_p M)^2}{R_{s1} + R_{eq}} \\ \dot{I}_{p1} = \frac{\dot{U}_{inv}}{Z_{in}} = \frac{\dot{U}_{inv} (R_{s1} + R_{eq})}{R_{p1} (R_{s1} + R_{eq}) + (\omega_p M)^2} \\ \dot{I}_{s1} = - \frac{j\omega_p M \dot{U}_{inv}}{R_{p1} (R_{s1} + R_{eq}) + (\omega_p M)^2} \end{cases} \quad (4)$$

Defining path gain as the ratio of the load voltage to the full bridge inverter output voltage, we derive expressions for path gain, input power, output power, and system efficiency as

$$\begin{cases} G_p = \frac{-R_{eq} \dot{I}_{s1}}{\dot{U}_{inv}} = j \frac{8\omega_p M R_L}{\pi^2 R_{p1} R_{s1} + 8R_{p1} R_L + \pi^2 (\omega_p M)^2} \\ P_{in} = U_{inv} I_{p1} = \frac{(R_{s1} + R_{eq}) U_{inv}^2}{R_{p1} (R_{s1} + R_{eq}) + (\omega_p M)^2} \\ P_{out} = I_{s1}^2 R_{eq} = \frac{\omega_p^2 M^2 U_{inv}^2 R_{eq}}{[R_{p1} (R_{s1} + R_{eq}) + (\omega_p M)^2]^2} \\ \eta = \frac{P_{out}}{P_{in}} = \frac{\omega_p^2 M^2 R_{eq}}{(R_{s1} + R_{eq}) [R_{p1} (R_{s1} + R_{eq}) + \omega_p^2 M^2]} \end{cases} \quad (5)$$

2.4 Information Transmission

The forward information path is shown in Fig. 5. When source AC₁ is active, wave trappers (a) and (b) exhibit high impedance at power carrier f_p , this is equivalent to an open circuit. At f_i , Z_{p1} and Z_{s1} exhibit high impedance and are considered open circuits, while the impedance of the wave trappers on the forward information path is zero and is

considered a short circuit. In addition, the backward information path impedances Z_{r1} and Z_{r2} exhibit high impedance and are considered open circuit. So the information signal is transmitted via main coupling coils L_p , L_s and parallel resonant branch to the forward receiving resistor R_l . Here, R_{f1} and R_{f2} are the internal resistances of the forward signal transmission path, and Z_{p1} and Z_{s1} are the equivalent impedances of the power transmission path.

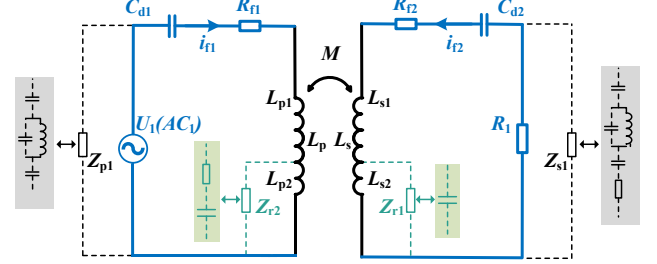


Fig. 5. Equivalent Schematic of Forward Information Transmission.

Under resonant operation, the forward information path satisfies

$$\begin{cases} \omega_1 = \frac{1}{\sqrt{L_p C_{d1}}} = \frac{1}{\sqrt{L_s C_{d2}}} \\ Z_{p1} = \frac{1}{j\omega C_{p1}} + \frac{1}{j\omega C_{t1c}} + \frac{j\omega L_{t1}}{1 - \omega^2 L_{t1} C_{t1}} \\ Z_{s1} = \frac{1}{j\omega C_{s1}} + \frac{1}{j\omega C_{t2c}} + \frac{j\omega L_{t2}}{1 - \omega^2 L_{t2} C_{t2}} + R_{eq} \end{cases} \quad (6)$$

According to Fig. 5, the KVL equations for primary and secondary loops are derived as

$$\begin{cases} \left(R_{f1} + j\omega L_p + \frac{1}{j\omega C_{d1}} \right) \dot{I}_{f1} + j\omega M \dot{I}_{f2} = \dot{U}_1 \\ \left(R_{f2} + j\omega L_s + \frac{1}{j\omega C_{d2}} \right) \dot{I}_{f2} + j\omega M \dot{I}_{f1} = \dot{U}_{Rl} \end{cases} \quad (7)$$

Where U_{Rl} is the voltage across the information receiving resistor R_l , and the forward information transmission path voltage gain G_f is defined as the ratio of U_{Rl} to the forward information source voltage U_1 . Therefore, at resonance, the forward information transmission gain G_f can be expressed as

$$G_f = \frac{\dot{U}_{Rl}}{\dot{U}_1} = \frac{j\omega_1 M R_l}{R_{f1} (R_{f2} + R_l) + (\omega_1 M)^2} \quad (8)$$

The backward information path follows the same frequency selection principle, but the carrier is injected through the partial inner coil, resulting in a different equivalent gain expression. Similarly, the equivalent circuit for backward information transmission satisfies

$$\omega_2 = \frac{1}{\sqrt{L_{p2}C_{p2}}} = \frac{1}{\sqrt{L_{s2}C_{s2}}} \quad (9)$$

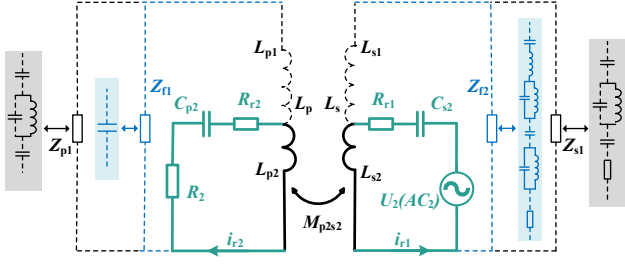


Fig. 6. Equivalent Schematic for Backward Information Transmission.

In Fig. 6, the equivalent impedances Z_{eqp} and Z_{eqs} of the power path and the forward information transmission path, which are connected in parallel with the equivalent inductances L_{p2} and L_{s2} of the partial coils, can be expressed as

$$\begin{cases} Z_{eqp} = j\omega L_{p1} + Z_{p1} \parallel Z_{f1} \\ Z_{eqs} = j\omega L_{s1} + Z_{s1} \parallel Z_{f2} \end{cases} \quad (10)$$

At f_2 , Z_{eqp} and Z_{eqs} exhibit high impedance and are equivalent to an open circuit; therefore, the backward information transmission path can be modeled as an SS type compensation topology. Defining the gain of the backward information transmission path as the ratio of the voltage U_{R2} across the information receiving resistor R_2 to U_2 , we can similarly derive the path gain G_r as

$$G_r = \frac{U_{R2}}{U_2} = \frac{j\omega_2 M_{p2s2} R_2}{R_{r1}(R_{r2} + R_2) + (\omega_2 M_{p2s2})^2} \quad (11)$$

Where M_{p2s2} represents the mutual inductance between partial coils, while R_{r1} and R_{r2} represent the internal resistances of the backward information transmission branches.

3 Design and Crosstalk Analysis

3.1 System Parameter Design

To simplify the theoretical calculations and derivations, symmetric parameter designs are employed for both the power transmission path and the information transmission channel, as described by (12). Furthermore, the coil parameters designed in this paper satisfy (13), where λ represents the coil partitioning ratio, which indicates the distribution ratio of the auxiliary power coils to the main coil within the system, in order to maintain high efficiency and low crosstalk performance.

$$\begin{cases} C_{d1} = C_{d2}, C_{p2} = C_{s2} \\ C_{p1} = C_{s1}, L_{p1} = L_{s1}, L_{p2} = L_{s2} \\ C_{t1c} = C_{t2c}, L_{t1} = L_{t2}, C_{t1} = C_{t2} \end{cases} \quad (12)$$

$$\begin{cases} L_{p1} = (1 - \lambda)L_p, L_{p2} = \lambda L_p \\ L_{s1} = (1 - \lambda)L_s, L_{s2} = \lambda L_s \end{cases} (0 < \lambda < 1) \quad (13)$$

During the parameter calculation process, the power transmission frequency f_p , the forward information carrier frequency f_1 , the backward information carrier frequency f_2 , and the inductance of the mutual inductance coil can be determined first; therefore, the corresponding resonance compensation capacitance values C_{p1} , C_{s1} , C_{d1} , C_{d2} , C_{p2} , and C_{s2} are

$$\begin{cases} C_{p1} = C_{s1} = \frac{1}{\omega_p^2 L_p} = \frac{1}{\omega_p^2 L_s} \\ C_{d1} = C_{d2} = \frac{1}{\omega_1^2 L_p} = \frac{1}{\omega_1^2 L_s} \\ C_{p2} = C_{s2} = \frac{1}{\omega_2^2 L_{p2}} = \frac{1}{\omega_2^2 L_{s2}} \end{cases} \quad (14)$$

As inductance increases, impedance increases as well; in a circuit, this results in better attenuation of the carrier at the center resonance frequency. When calculating parameters, one can first determine the inductance value and then calculate the capacitance value based on that inductance value; therefore

$$\begin{cases} C_{t1} = C_{t2} = \frac{1}{\omega_1^2 L_{t1}} = \frac{1}{\omega_1^2 L_{t2}} \\ C_{t3} = \frac{1}{\omega_p^2 L_{t3}}, C_{t4} = \frac{1}{\omega_2^2 L_{t4}} \end{cases} \quad (15)$$

To prevent the addition of a wave trapper from affecting the system's original operating conditions, appropriate compensation capacitors or inductors must be connected in series. Taking wave trappers (a) and (c) as examples, C_{t1c} and L_{t3c} should satisfy

$$\begin{cases} -j \frac{1}{\omega_p C_{t1c}} + \frac{j\omega_p L_{t1}}{1 - \omega_p^2 L_{t1} C_{t1}} = 0 \\ j\omega_1 L_{t3c} + \frac{j\omega_1 L_{t3}}{1 - \omega_1^2 L_{t3} C_{t3}} = 0 \end{cases} \quad (16)$$

Based on (16), it follows that C_{t1c} , C_{t2c} , C_{t4c} , and L_{t3c} can be expressed as

$$\begin{cases} C_{t1c} = C_{t2c} = \frac{1 - \omega_p^2 L_{t1} C_{t1}}{\omega_p^2 L_{t1}} = \frac{1 - \omega_p^2 L_{t2} C_{t2}}{\omega_p^2 L_{t2}} \\ C_{t4c} = \frac{1 - \omega_1^2 L_{t4} C_{t4}}{\omega_1^2 L_{t4}}, L_{t3c} = \frac{L_{t3}}{\omega_1^2 L_{t3} C_{t3} - 1} \end{cases} \quad (17)$$

To reduce the number of components used, some components can be simplified into a single component using impedance series or parallel formulas, thereby simplifying the system.

3.2 Coil Design

Since circular coils produce a more uniform magnetic

field distribution, this paper employs a circular planar helical coil as the coupling mechanism; a schematic diagram of the coil structure is shown in Fig. 7.

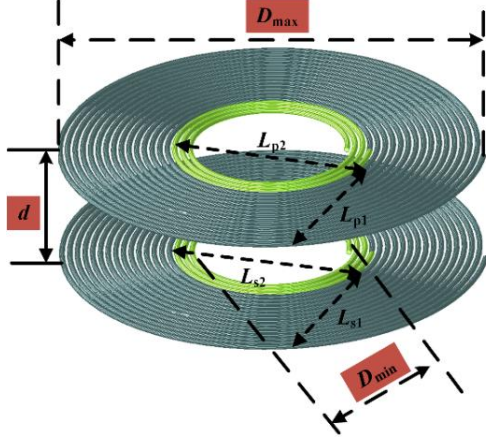


Fig. 7. Partial Coil Structure in 3D.

In Fig. 7, $D_{\max}$ denotes the outer diameter of the coil, and $D_{\min}$ denotes the inner diameter of the coil. For tightly wound planar helical coils, the inductance can be calculated using an empirical formula, expressed as

$$L = \frac{\mu_0 N^2 D_a}{2} \left[\ln \left(\frac{2.46}{\beta} \right) + 0.2 \beta^2 \right] \quad (18)$$

Here, L represents the self inductance of the coil, N represents the number of turns, $D_a = (D_{\max} + D_{\min})/2$ represents the equivalent average diameter of the coil, $\beta = (D_{\max} - D_{\min})/(D_{\max} + D_{\min})$, and μ_0 is the magnetic permeability of free space. From this, it can be seen that the inductance of the coil is primarily influenced by the number of turns and the inner and outer diameters.

Therefore, when two coaxial coils with symmetrical parameters are separated by a distance d , their mutual inductance M and coupling coefficient k can be expressed as

$$\begin{cases} M = \frac{\mu_0 \pi N^2 D_a^4}{2(d^2 + D_a^2)^{3/2}} \\ k = \frac{M}{L} = \frac{\pi D_a^3}{(d^2 + D_a^2)^{3/2} \left[\ln \left(\frac{2.46}{\beta} \right) + 0.2 \beta^2 \right]} \end{cases} \quad (19)$$

It can be observed that the coupling coefficient k is negatively correlated with transmission distance d , and k decreases monotonically with the increase of d . Increasing the equivalent coil diameter D_a reduces the proportion of d in the denominator term. In other words, a larger D_a leads to a slower variation rate of k against d , which makes the coupling characteristic less sensitive to distance fluctuation.

In practical applications, axial distance fluctuation and

lateral misalignment between the transmitting and receiving coils are inevitable. Accordingly, the coil dimension can be properly enlarged within available installation space to improve the stability of coupling coefficient k against spatial position variation.

According to the coil configuration illustrated in Fig. 7, a finite element model of coupled coils is constructed on the Ansys Maxwell platform to simulate electromagnetic field propagation characteristics in free space. Copper with high electrical conductivity is defined as the coil material, and the corresponding simulation results are presented in Fig. 8.

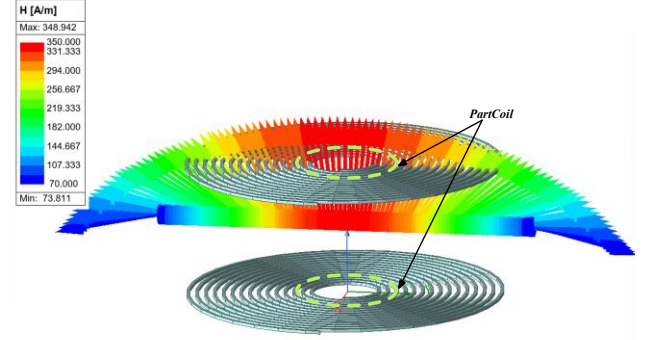


Fig. 8. Scalar Plot of the Magnetic Field Distribution.

As shown in Fig. 8, the magnetic field intensity vector reaches its peak near the inner turns of the coil and decays monotonically outward, which is consistent with typical near-field coupling characteristics. Therefore, the inner region of the coil is selected for information transmission..

Define k_{ps} and k_{p2s2} as the coupling coefficients between full coils L_p , L_s and partial inner coils L_{p2} , L_{s2} respectively. $k_{ps} \approx k_{p2s2}$ holds under the ideal uniform magnetic field condition. According to the field distribution in Fig. 8, the practical magnetic field is non-uniform. Owing to the fewer turns of the inner partial coils, therefore, in practice $k_{ps} \leq k_{p2s2}$, so M_{p2s2} can be expressed as

$$M_{p2s2} = k_{p2s2} \sqrt{L_{p2} L_{s2}} = \lambda \frac{k_{p2s2}}{k_{ps}} M \quad (20)$$

In addition, the skin effect is considered in high-frequency coil design. High-frequency alternating current concentrates on conductor surface, reducing inner current and raising equivalent resistance. Governed by operating frequency and conductor material, the skin depth is defined as

$$\delta = \sqrt{\frac{2}{\omega \mu \sigma}} \quad (21)$$

Where ω is the angular frequency, μ is the magnetic permeability of the conductor material, and σ is the electrical conductivity.

3.3 Mutual Interference Between Power and Information

3.3.1 Analysis of Crosstalk From Information Transmission on Power Transmission

The interference imposed by information channel on power transmission is analyzed from two aspects: whether the access of information branch brings extra power loss, and whether the operating signal of information source disturbs power delivery. Combined with Fig.5, (4) and (15), it is derived that

$$\begin{cases} Z_{dp2} = Z_{r2} \parallel j\omega L_{p2} = \frac{j\omega_p L_{p2}(1 + j\omega_p R_2 C_{p2})}{1 + j\omega_p R_2 C_{p2} - \omega_p^2 / \omega_2^2} \\ Z_{ds2} = Z_{r1} \parallel j\omega L_{s2} = \frac{j\omega_p L_{s2}}{1 - \omega_p^2 / \omega_2^2} \end{cases} \quad (22)$$

Where Z_{dp2} and Z_{ds2} are the equivalent impedances obtained by parallel connection of backward information branch and partial coil inductance on primary and secondary sides, respectively.

In this system, the forward and backward information carrier frequencies are set much higher than the power carrier frequency ($f_2 > f_1 \gg f_p$), hence $1 - \omega_p^2 / \omega_2^2 \approx 1$, so (22) can be written as

$$\begin{cases} Z_{dp2} \approx j\omega_p L_{p2} = Z_{ppe} \\ Z_{ds2} \approx j\omega_p L_{s2} = Z_{spe} \end{cases} \quad (23)$$

Where Z_{ppe} and Z_{spe} denote the impedances of partial coils at power transmission frequency. At power transmission frequency, the impedance of backward information branch is much larger than that of partial coils, so the backward information branch can be regarded as open circuit when paralleled. Accordingly, the parallel impedance Z_{dp1} between forward information branch and primary coil inductance L_p is expressed as

$$Z_{dp1} = Z_{f1} \parallel j\omega L_p = \frac{j\omega_p L_p}{1 - \omega_p^2 / \omega_1^2} \approx j\omega_p L_p = Z_{Lp} \quad (24)$$

Where Z_{Lp} denotes the primary coil impedance at power transmission frequency. Define the calculation error of the above approximation as α_1 , which is expressed as

$$\alpha_1 = \left| \frac{|Z_{dp1}| - |Z_{Lp}|}{|Z_{dp1}|} \right| = \left| 1 - \frac{\omega_p^2}{\omega_1^2} \right| \quad (25)$$

The equivalent error curve is plotted based on (25), as shown in Fig. 9.

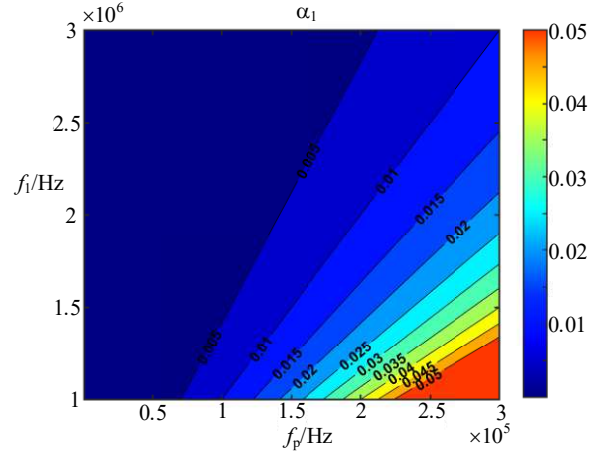


Fig. 9. Error α_1 Variation Curve.

The results indicate that when the power transmission frequency f_p is below 100 kHz and the forward information carrier frequency f_1 is above 1 MHz, the equivalent error α_1 is less than 1% and can be neglected. This demonstrates that the approximation of treating the forward information transmission branch as an open circuit is valid under the frequency conditions set in this paper ($f_2 > f_1 \gg f_p$). Thus, during power transmission, the signal path impedance is extremely high, yielding nearly zero additional power loss.

3.3.2 Analysis of Crosstalk From Power Transmission on Information Transmission

In practical engineering, the wave trapper cannot achieve complete theoretical blocking, so it is necessary to analyze the interference of power transmission on the information link. The interference from power transmission is mainly the low frequency interference voltage generated by the power carrier on the information receiving resistor, and its equivalent interference model is shown in Fig. 10.

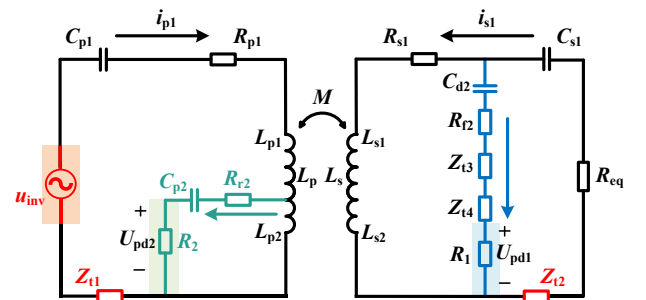


Fig. 10. Equivalent Circuit of Power to Information Interference.

In Fig. 10, U_{pd1} and U_{pd2} are the interference voltages on information receiving resistors R_1 and R_2 , respectively. Let the terminal voltages of primary and secondary coils be U_{Lp} and U_{Ls} , and the voltage across inductor U_{Lp2} be U_{Ls2} . The influence of information branch can be neglected when calculating the voltages of power loop. Define G_{p_Lp2} and G_{p_Ls} as the voltage transfer gains from inverter voltage source u_{inv} to U_{Lp2} and U_{Ls} , G_{Lp2_R2} and G_{Ls_R1} as the voltage transfer gains from U_{Lp2} to U_{pd2} and from U_{Ls} to U_{pd1} ,

respectively. Combined with (3), (4) and (13), it is derived that

$$\begin{cases} G_{p_Lp2} = \frac{\dot{U}_{Lp2}}{\dot{U}_{inv}} = \frac{\lambda \dot{U}_{Lp}}{\dot{U}_{inv}} = \frac{(R_{p1} + j\omega_p L_p) \lambda}{R_{p1} + Z_{t1} + \frac{(\omega_p M)^2}{R_{s1} + R_{eq} + Z_{t2}}} \\ G_{p_Ls} = \frac{\dot{U}_{Ls}}{\dot{U}_{inv}} = \frac{j\omega_p M \cdot \frac{R_{s1} + j\omega_p L_s}{R_{p1} + Z_{t1}}}{R_{s1} + R_{eq} + Z_{t2} + \frac{(\omega_p M)^2}{R_{p1} + Z_{t1}}} \\ G_{Lp2_R2} = \frac{\dot{U}_{pd2}}{\dot{U}_{Lp2}} = \frac{R_2}{R_2 + R_{r2} + \frac{1}{j\omega_p C_{p2}}} \\ G_{Ls_R1} = \frac{\dot{U}_{pd1}}{\dot{U}_{Ls}} = \frac{R_1}{R_1 + R_{f2} + Z_{t3} + Z_{t4} + \frac{1}{j\omega_p C_{d2}}} \end{cases} \quad (26)$$

Where Z_{t3} and Z_{t4} are the equivalent impedances of wave trappers (c) and (d), respectively, which satisfy

$$\begin{cases} Z_{t3} = j\omega L_{t3c} + \frac{j\omega L_{t3}}{1 - \omega^2 L_{t3} C_{t3}} \\ Z_{t4} = \frac{1}{j\omega C_{t4c}} + \frac{j\omega L_{t4}}{1 - \omega^2 L_{t4} C_{t4}} \end{cases} \quad (27)$$

Combined with (26) and (27), the interference voltage gains of power transmission on forward and backward information links are obtained as

$$\begin{cases} G_{p_R1} = \frac{\dot{U}_{pd1}}{\dot{U}_{inv}} = \frac{\dot{U}_{Ls}}{\dot{U}_{inv}} \cdot \frac{\dot{U}_{pd1}}{\dot{U}_{Ls}} = G_{p_Ls} G_{Ls_R1} \\ G_{p_R2} = \frac{\dot{U}_{pd2}}{\dot{U}_{inv}} = \frac{\dot{U}_{Lp2}}{\dot{U}_{inv}} \cdot \frac{\dot{U}_{pd2}}{\dot{U}_{Lp2}} = G_{p_Lp2} G_{Lp2_R2} \end{cases} \quad (28)$$

The above analysis shows that when the system parameters are determined, the interference gain G_{p_R1} is related to the information receiving resistor R_1 , equivalent load resistance R_{eq} and mutual inductance M . The interference gain G_{p_R2} is related to R_2 , R_{eq} , M and the coil turns ratio λ . Accordingly, the interference voltages can be expressed

$$\begin{cases} \dot{U}_{pd1} = \dot{U}_{inv} G_{p_Ls} G_{Ls_R1} \\ \dot{U}_{pd2} = \dot{U}_{inv} G_{p_Lp2} G_{Lp2_R2} \end{cases} \quad (29)$$

Let SNR_1 , SNR_2 be the SNRs of forward and backward information paths, respectively, Combined with (8), (11)

and (29), SNR_1 , SNR_2 can be regarded as

$$\begin{cases} SNR_1 = 20 \lg \frac{U_{R1}}{U_{pd1}} = 20 \lg \frac{U_1 G_f}{U_{inv} G_{p_R1}} \\ SNR_2 = 20 \lg \frac{U_{R2}}{U_{pd2}} = 20 \lg \frac{U_2 G_r}{U_{inv} G_{p_R2}} \end{cases} \quad (30)$$

Based on the above analysis, it can be seen that increasing the information transmission gains G_f and G_r , or reducing the interference voltage gains G_{p_R1} and G_{p_R2} both can improve the quality of information transmission.

3.4 Crosstalk Between Information Paths

In addition to power induced interference, mutual interference also exists between information signals. Such interference mainly appears as interference voltage from a same side signal source across the local receiving resistor. Taking the influence of forward information transmission on the backward information path as an example, this phenomenon is illustrated in Fig. 11.

Due to mutual coupling among coils, the mutual inductance between any two coils is illustrated above. Since L_{p1} and L_{p2} are series coupled, $L_p = L_{p1} + L_{p2} + 2M_{p1p2}$. Similarly, $L_s = L_{s1} + L_{s2} + 2M_{s1s2}$. Therefore, based on KVL and KCL, the equations in Fig. 11 can be derived as follows

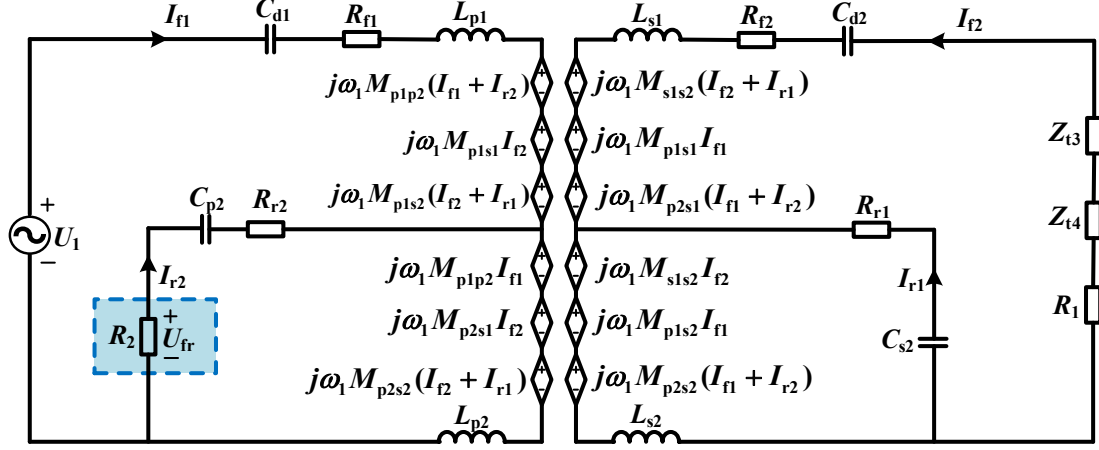


Fig. 11. Equivalent Circuit from Forward Information Transmission Interference on Backward Information

$$\begin{bmatrix} \dot{U}_1 \\ 0 \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} R_{f1} & j\omega_1 M \\ j\omega_1 M & R_1 + R_{f2} \\ j\omega_1 M_{eq1} & j\omega_1 (M_{s1s2} + L_{s2}) \\ j\omega_1 (M_{p1p2} + L_{p2}) & j\omega_1 M_{eq2} \end{bmatrix} \begin{bmatrix} \dot{I}_{f1} \\ \dot{I}_{f2} \\ \dot{I}_{r1} \\ \dot{I}_{r2} \end{bmatrix} \quad (31)$$

Let (31) be rewritten as $u_l = Z_l i$, where Z_l denotes the transformation matrix. We then obtain $u_l = [U_1 \ 0 \ 0 \ 0]^T$, $i = [I_{f1} \ I_{f2} \ I_{r1} \ I_{r2}]^T$, therefore, the current vector can be expressed as $i = Z_l^{-1} u_l$, in which M , M_{eq1} , M_{eq2} , Z_{eqp2} , and Z_{eqs2} are

$$\begin{cases} Z_{eqp2} = Z_{eqs2} = \frac{1 - \omega_1^2 L_{p2} C_{p2}}{j\omega_1 C_{p2}} = \frac{1 - \omega_1^2 L_{s2} C_{s2}}{j\omega_1 C_{s2}} \\ M_{eq1} = M_{p1s1} + M_{p2s2}, \quad M_{eq2} = M_{p2s1} + M_{p2s2} \\ M = M_{p1s1} + M_{p1s2} + M_{p2s1} + M_{p2s2} \end{cases} \quad (32)$$

The interference gain G_{f-R2} from the forward input source U_1 to the voltage U_{fr} across resistor R_2 can be expressed as

$$G_{f-R2} = \frac{U_{fr}}{U_1} = \frac{-I_{r2} R_2}{U_1}$$

Similarly, analyze the interference from backward information transmission on the forward information, namely the interference voltage U_{rf} across the forward receiving resistor R_1 when the backward source U_2 is active. The equivalent circuit is shown in Fig. 12, where

$$\begin{bmatrix} 0 \\ 0 \\ \dot{U}_2 \\ 0 \end{bmatrix} = \begin{bmatrix} R_{f1} + Z_{eqd1} & j\omega_2 M \\ j\omega_2 M & R_{f2} + Z_{eqd2} \\ j\omega_2 M_{eq1} & j\omega_2 (M_{s1s2} + L_{s2}) \\ j\omega_2 (M_{p1p2} + L_{p2}) & j\omega_2 M_{eq2} \end{bmatrix} \begin{bmatrix} \dot{I}_{f1} \\ \dot{I}_{f2} \\ \dot{I}_{r1} \\ \dot{I}_{r2} \end{bmatrix} \quad (34)$$

As mentioned earlier, let (34) be rewritten as $u_2 = Z_2 i$, where Z_2 denotes the transformation matrix. We then obtain $u_2 = [0 \ 0 \ U_2 \ 0]^T$, i remains constant, so the current vector can be expressed as $i = Z_2^{-1} u_2$, in which Z_{eqd1} , Z_{eqd2} are

$$\begin{cases} Z_{eqd1} = \frac{1 - \omega_2^2 L_p C_{d1}}{j\omega_2 C_{d1}} \\ Z_{eqd2} = Z_{t3} + Z_{t4} + R_1 + \frac{1 - \omega_2^2 L_s C_{d2}}{j\omega_2 C_{d2}} \end{cases} \quad (35)$$

Therefore, the interference gain G_{r-R1} of the voltage U_{rf} across the receiving resistor R_1 from the information source U_2 can be expressed as

$$G_{r-R1} = \frac{U_{rf}}{U_2} = \frac{-I_{f2} R_1}{U_2} \quad (36)$$

In summary, information crosstalk has two main paths: direct coupling between coils on the same side (e.g., M_{p1p2} , M_{s1s2}) and secondary coupling via multiple coil paths. The small tap ratio of partial coils results in very weak same-side coupling, and the secondary path introduces large attenuation. Thus, inter information crosstalk is very low, which is a key advantage of partial coils.

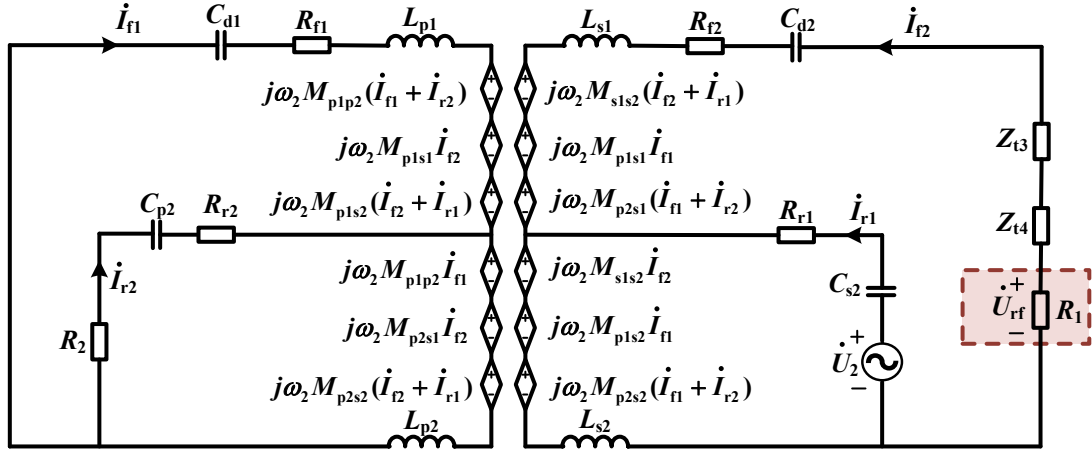


Fig. 11. Equivalent Circuit from Backward Information Transmission Interference on Forward Information

4 Simulation Analysis and Experimental Results

Based on the theoretical derivations presented above and the simulation parameters listed in Table 1, we conducted quantitative simulations to analyze the system's power transfer characteristics, crosstalk characteristics, and coil structure characteristics. This analysis clarifies how various parameters affect system performance, thereby providing a theoretical basis for parameter tuning and performance optimization of the experimental platform.

Table 1 Parameters Used in the Simulation

Parameters	Value	Parameters	Value
E_{dc}	30 V	f_p	100 kHz
L_p, L_s	100 μ H	C_{p1}, C_{s1}	25.33 nF
f_1	1 MHz	C_{d1}, C_{d2}	253.3 pF
R_{p1}, R_{p2}	0.3 Ω	L_{t1}, L_{t2}	300 μ H
C_{t1}, C_{t2}	84.43 pF	C_{t1e}, C_{t2e}	8.35 nF
f_2	3 MHz	C_{t1}, C_{t2}	84.43 pF
L_{t3}	220 μ H	L_{t4}	22 μ H
C_{t3}	11.5 nF	C_{t4}	127 pF
R_{f1}, R_{f2}	0.2 Ω	R_{t1}, R_{t2}	0.15 Ω

4.1 Power Transfer Characteristics Curve

After determining the system power transmission frequency, coil inductance and internal resistance parameters, the power characteristics of the system can be analyzed according to (2), (5) and parameters of Table 1, and the relationship curve between output power P_{out} , coil mutual inductance M and equivalent load R_{eq} is plotted, as shown in Fig. 12.

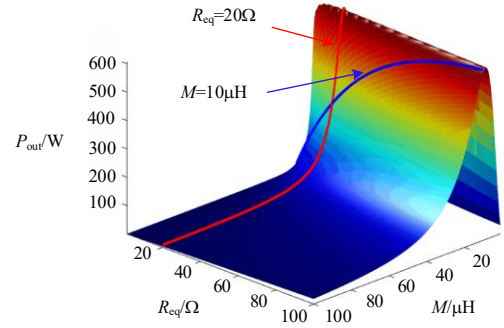


Fig. 12. Curve of P_{out} as a Function of M and R_{eq} .

It can be observed from Fig. 12. that the system output power P_{out} exhibits a trend of first increasing and then decreasing with mutual inductance M . At low mutual inductance, the weak coupling between primary and secondary sides limits the energy transferred to the secondary side, resulting in low output power. As mutual inductance increases, the enhanced coil coupling improves the system transmission capability, and the output power rises gradually. When mutual inductance exceeds the optimal value, the increased secondary reflected impedance reduces the primary current, causing the system to deviate from the optimal impedance matching state and the output power to decrease.

Meanwhile, the output power P_{out} is also affected by the equivalent load R_{eq} . At a fixed mutual inductance, there exists an optimal load value that maximizes the output power. When the load deviates from this value, the impedance matching fails and the output power decreases. Since the denominator of P_{out} in (5) is mainly determined by the product of mutual inductance and angular frequency, P_{out} can be approximately considered proportional to R_{eq} in engineering applications. Similarly, the variation curve of system transmission efficiency η with mutual inductance M and equivalent load R_{eq} is plotted according to (5), as shown

in Fig. 13.

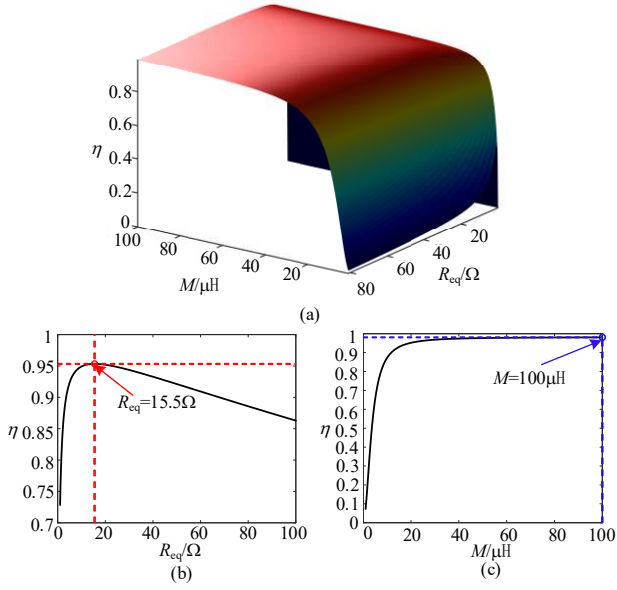


Fig. 13. Graph of the η as a function of M and R_{eq} : (a) 3D plot of η as a function of M and R_{eq} ; (b) η varies with R_{eq} ; (c) M varies with R_{eq} .

It can be seen from Fig. 13. that the system transmission efficiency η generally increases with mutual inductance M . At low mutual inductance, the weak coupling between primary and secondary sides results in low system efficiency. As mutual inductance increases, the enhanced coil coupling transfers more energy to the load, and the efficiency rises gradually. When mutual inductance exceeds $20 \mu\text{H}$, the growth rate tends to flatten out. The efficiency η exhibits a trend of first increasing and then decreasing with equivalent load R_{eq} : when R_{eq} is small, the limited power absorption capability of the load leads to low efficiency; as R_{eq} increases, the system approaches the optimal matching state and the efficiency rises; when R_{eq} exceeds the optimal value, the efficiency decreases accordingly.

4.2 Power and Information Crosstalk Characteristics Curve

Combined with (8) and Table 1, the variation curve of forward channel gain G_f with coil mutual inductance M and sampling resistor R_l is plotted, as shown in Fig. 14.

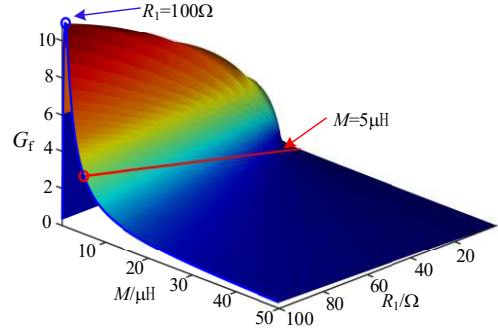


Fig. 14. Function curves of G_f with respect to M and R_l

It can be observed from Fig. 14. that the forward information transmission channel gain G_f exhibits a trend of first increasing and then decreasing with mutual inductance M . When M larger than $20 \mu\text{H}$, the channel gain remains low even with a large sampling resistor, so the mutual inductance M should be kept below $20 \mu\text{H}$ in design. G_f increases rapidly first and then tends to flatten out with the increase of sampling resistor R_l , so the value of R_l can be appropriately increased in engineering design.

Similarly, the variation curve of backward channel gain G_r with mutual inductance M_{p2s2} and sampling resistor R_2 is plotted according to the parameters given in Table 1, as shown in Fig. 15.

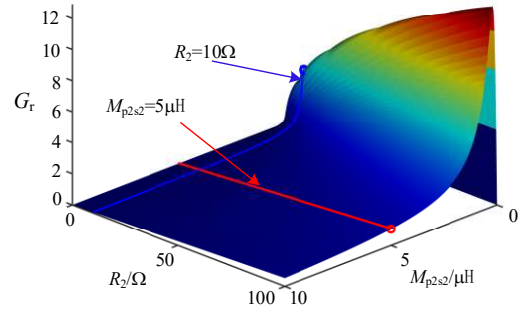


Fig. 15. Function curves of G_r with respect to M_{p2s2} and R_2

It can be seen from Fig. 15. that the backward information transmission channel gain G_r exhibits a trend of first increasing and then decreasing with mutual inductance M_{p2s2} . When M_{p2s2} larger than $5 \mu\text{H}$, the channel gain remains low even with a large sampling resistor R_2 , so M_{p2s2} should be controlled below $5 \mu\text{H}$ in design. Since the denominator in (11) is mainly determined by the product of M_{p2s2} and angular frequency ω_2 , and R_2 accounts for a negligible proportion in the denominator, the backward channel gain G_r can be approximately considered positively correlated with R_2 . However, R_2 is not the larger the better,

as its increase will lead to a synchronous rise in interference voltage. Therefore, R_2 should be no less than 10Ω in design.

4.3 Gain Variation Curve of the Information Transmission Path

Combined with (28) and the parameters given in Table 1, the variation curve of forward interference gain G_{p_R1} with mutual inductance M , sampling resistor R_1 and equivalent load R_{eq} is obtained, as shown in Fig. 16.

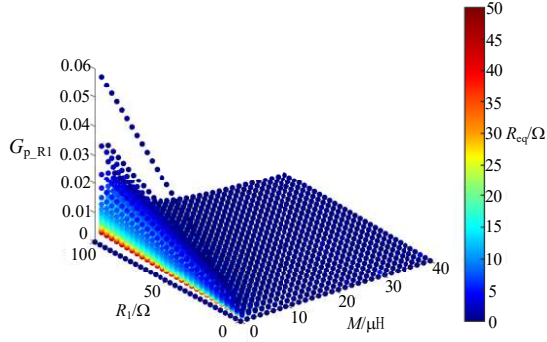


Fig. 16. Variation of G_{p_R1} with respect to M , R_1 , and R_{eq}

It can be seen from Fig. 16 that the forward interference gain G_{p_R1} is positively correlated with the sampling resistor R_1 , and exhibits a trend of first increasing and then decreasing with mutual inductance M , which is consistent with the variation of output power with M . The higher the transmission power, the more energy transferred to the secondary side, and the stronger the interference. Meanwhile, G_{p_R1} is negatively correlated with the equivalent load R_{eq} , so R_{eq} should be no less than 10Ω in design.

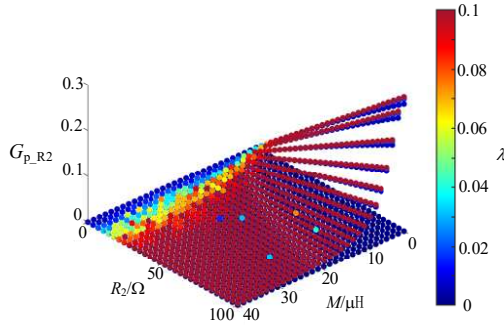


Fig. 17. Variation of G_{p_R2} with respect to M , R_2 , and λ at $R_{eq} = 20 \Omega$

Considering that the backward interference gain G_{p_R2} is related to four variables, dimensionality reduction analysis is required first. It can be seen from (29) that G_{p_R2} is positively correlated with the equivalent load R_{eq} , so the value of R_{eq} should be reduced. Combined with the system efficiency and power characteristic curves, $R_{eq} = 20 \Omega$ is selected for analysis, and the variation curve of G_{p_R2} with relevant parameters is obtained, as shown in Fig. 17.

It can be seen from Fig. 17 that the backward interference gain G_{p_R2} is positively correlated with the sampling resistor

R_2 and the coil turns ratio λ , and approximately negatively correlated with the mutual inductance M . This is because within a certain range, the increase of M enhances the energy transferred to the secondary side, which reduces the primary side voltage U_{Lp2} and thus the interference to the backward information. To suppress the interference, the coil turns ratio λ should be kept below 0.08 in design.

4.4 Experimental Results

The actual measured values for the system parameters are shown in Table 2.

Table 2 Actual Measured System Parameters

Parameters	Value	Parameters	Value
E_{dc}	30V	f_p	100kHz
L_p, L_s	99.6/99.7μH	λ	0.05
f_1, f_2	1/3MHz	f_2	3MHz
L_{p2}, L_{s2}	4.9/4.8μH	M	19.7μH
C_{d1}	253.4pF	C_{d2}	253.3pF
C_{t1c}, C_{t2c}	8.3/8.4nF	R_L	20~50Ω
C_{t3}	11.5nF	C_{t4}	126.7nF
M_{p2s2}	0.85μH	R_1, R_2	120/20Ω
C_{p1}, C_{s1}	25.3/25.7nF	C_{t1}, C_{t2}	84.2/84.4pF
R_{p1}	73.9mΩ	R_{s1}	73.4mΩ
L_{t3c}	2.22μH	C_{t4c}	1.02nF
L_{t3}	218.7μH	L_{t4}	22.5μH
C_{p2}	562.2pF	C_{s2}	561.9pF4

The experimental platform shown in Fig. 18 was built to verify the system and analysis. It consists of a DC supply, full bridge inverter, MCU, auxiliary supply, oscilloscope, main circuit (including modulation and demodulation), rectifier-filter, variable load and coupling coils.

Its operation includes power and information transmission. For power transfer, the full bridge inverter converts DC to AC and feeds it to the main circuit. Power is transferred via coupling coils from primary to secondary, then rectified and filtered to supply the load. For communication, forward and backward signals are transmitted through separate branches and resonant networks via coupling coils to the receiving resistors, then demodulated. This realizes non interfering power and full-duplex information transmission.

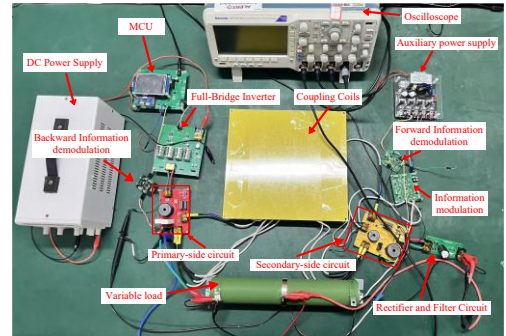


Fig. 18. Experimental Platform.

ASK modulation is used for its simplicity, easy demodulation and low cost. DDS (AD9833) generates the sinusoidal carrier, and an analog switch (ADG1419) driven

by binary control signals realizes amplitude modulation.

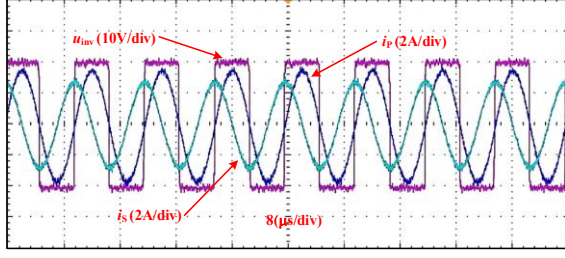


Fig. 19. Inverter Output Waveforms.

First, experiments were conducted at $E_{dc} = 30$ V, 100 kHz, and $R_L = 20 \Omega$. As shown in Fig. 19, the primary voltage (u_{inv}) and currents (i_p , i_s) exhibit minor high-frequency spikes but low overall harmonic distortion, indicating normal power transfer.

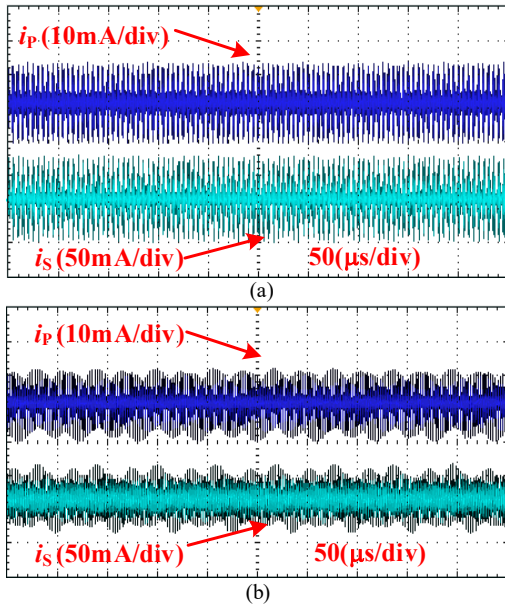


Fig. 20. Current waveforms during full duplex communication only: (a) Communication rate of 100 kbit/s; (b) Communication rate of 200 kbit/s.

Next, the impact of full duplex communication on power transfer was verified with the DC source turned off. As shown in Fig. 20, the measured maximum primary and secondary currents at data rates of 100 kbit/s and 200 kbit/s are 13.1 mA and 63.6 mA, respectively, indicating that the effect of information transmission on power delivery is negligible.

With the DC voltage maintained at 30 V and the full duplex communication rate set to 200 kbit/s, the system output power and transmission efficiency were tested under varying load resistances. The results are presented in Fig. 21.

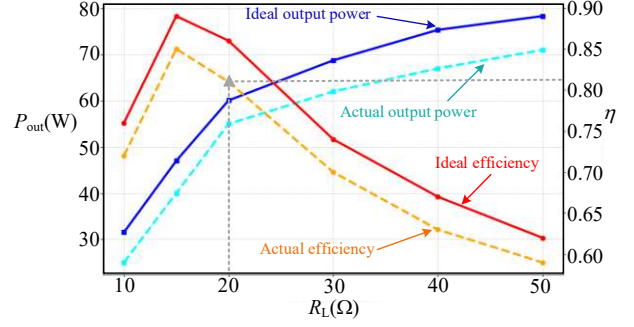


Fig. 21. P_{out} and η under varying R_L .

Experimental results show that with a 20 Ω load, the system output power is 54.7 W (approximately 5.3 W lower than the ideal value), and the transmission efficiency is 82.1% (a decrease of about 4.6%). This degradation is primarily attributed to conduction and switching losses in the inverter and rectifier, as well as additional power dissipation caused by the internal resistance of the blocking inductor.

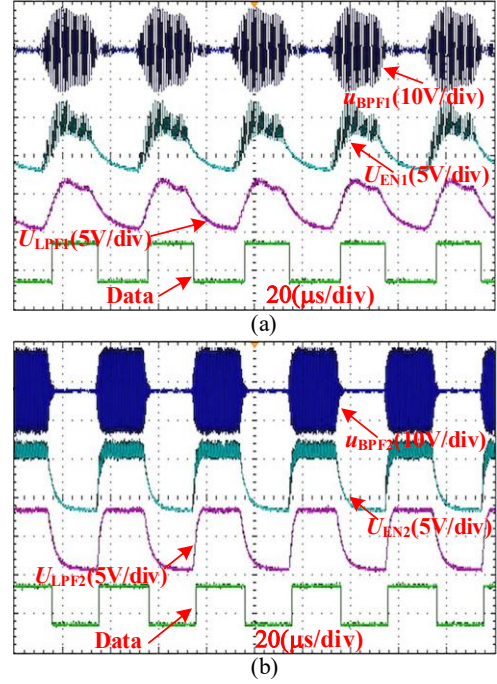


Fig. 22. Information demodulation waveforms at 50 kbit/s: (a) Forward information transmission; (b) Backward information transmission.

With the DC voltage set to 30 V and the load resistance at 20 Ω , the relevant waveforms measured at a fullduplex communication rate of 50 kbit/s are presented in Fig. 22.

In Fig. 22, u_{BPF1} and u_{BPF2} represent the bandpass filter output voltages, U_{EN1} and U_{EN2} denote the envelope detector output voltages, and U_{LPF1} and U_{LPF2} indicate the low-pass filter output voltages. It can be observed that full-duplex communication operates normally, with no bit errors occurring during demodulation at a transmission rate of 50 kbit/s.

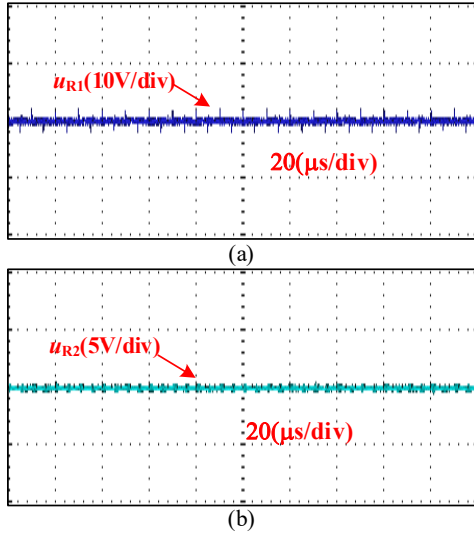


Fig. 23. Information sampling voltage waveforms during power transfer only: (a) Forward information sampling voltage U_{R1} ; (b) Backward information sampling voltage U_{R2} .

As shown in Fig. 23, the interference voltage caused by power transfer is very low. Furthermore, since the sampled voltage undergoes subsequent bandpass filtering, the impact of power on information demodulation can be neglected.

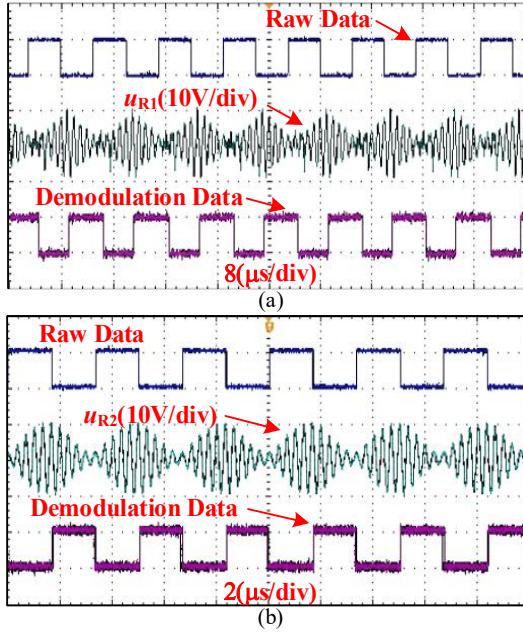


Fig. 24. Full-duplex information waveforms at different rates: (a) Forward communication rate of 200 kbit/s; (b) Backward communication rate of 600 kbit/s.

In practice, as the communication rate increases, the amplitude envelope of the information sampling voltage gradually decays. To ensure transmission quality and demodulation accuracy, this paper requires each bit to contain at least five carrier peaks. This guarantees sufficient dynamic response capability for envelope detection,

thereby avoiding decision errors between "1" and "0" signals. Fig. 24 illustrates the corresponding waveforms for a forward communication rate of 200 kbit/s and a backward communication rate of 600 kbit/s. The results demonstrate correct information demodulation, indicating that the maximum achievable communication rates for the forward and reverse directions are 200 kbit/s and 600 kbit/s, respectively.

Based on measured power and communication performance, the proposed system is compared with typical SWPIT schemes.

Most existing studies involve trade-offs between efficiency, data rate, communication mode and topology complexity. In contrast, the proposed scheme achieves high performance simultaneously in three key aspects: power transfer efficiency, data rate and communication scheme. The comparison is shown in Fig. 25.

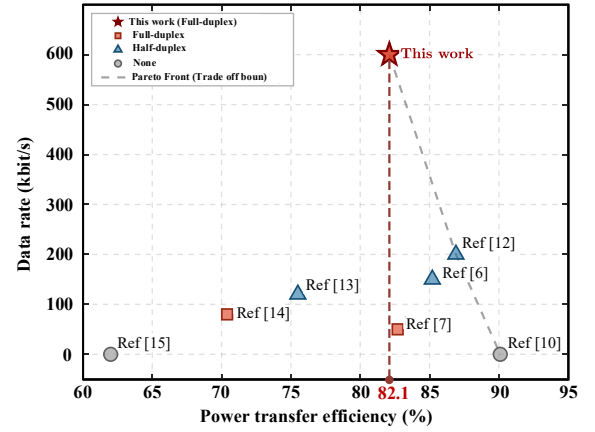


Fig. 25. Performance Comparison.

As shown in Fig. 25, the proposed scheme achieves 82.1 % power transfer efficiency under experimental conditions with full duplex communication. This efficiency is comparable to or higher than most existing SWPIT systems. Other methods often achieve high efficiency only with limited communication, or boost data rates at the cost of much higher circuit complexity.

Besides, the system supports bidirectional communication at 200 kbit/s forward and 600 kbit/s backward. Although its efficiency is not the absolute highest among reported works, it achieves the highest experimentally verified data rate among compared full duplex SWPIT systems, while retaining competitive efficiency. It thus balances power transfer and communication performance well.

These quantitative results confirm that the proposed scheme effectively relieves the traditional trade offs among efficiency, data rate and system complexity, achieving more balanced and comprehensive performance.

5 Conclusions

This paper proposes a full duplex SWPIT system using wave trappers and a partial power coil. Using their frequency selective features, the system physically isolates the power path from forward and backward information paths, greatly reducing crosstalk. ASK modulation simplifies the design while balancing data rate and power efficiency. Theoretical and experimental results verify stable, efficient full duplex operation at 54.7 W, with 200 kbit/s forward and 600 kbit/s backward data rates. Tests also show reliable power performance and strong interference suppression under the tested operating conditions. Compared with existing full duplex SWPIT systems, the proposed scheme features simpler structure, easy integration, low cost and high scalability, offering a practical solution for efficient full duplex communication between wireless power supplies and smart terminals.

References

- [1] Hui, S. Y. R., Zhong, W., Lee, C. K.: A Critical Review of Recent Progress in Mid-Range Wireless Power Transfer. *IEEE Trans. Power Electron.* 29(9), 4500–4511 (2014)
- [2] Zhang, Z., Pang, H., Georgiadis, A., Cecati, C.: Wireless Power Transfer—An Overview. *IEEE Trans. Ind. Electron.* 66(2), 1044–1058 (2019)
- [3] Trigui, A., Hached, S., Ammari, A. C., Savaria, Y., Sawan, M.: Maximizing Data Transmission Rate for Implantable Devices Over a Single Inductive Link: Methodological Review. *IEEE Rev. Biomed. Eng.* 12, 72–87 (2019)
- [4] Lu, X., Niyato, D., Wang, P., Kim, D. I.: Wireless Charger Networking for Mobile Devices: Fundamentals, Standards, and Applications. *IEEE Wireless Commun.* 22(2), 126–135 (2015)
- [5] Yan, Z., Song, B., Zhang, Y., Zhang, K., Mao, Z., Hu, Y.: A Rotation-Free Wireless Power Transfer System With Stable Output Power and Efficiency for Autonomous Underwater Vehicles. *IEEE Trans. Power Electron.* 34(5), 4005–4008 (2019)
- [6] Li, Z., Zhu, C., Jiang, J., Song, K., Wei, G.: A 3-kW Wireless Power Transfer System for Sightseeing Car Supercapacitor Charge. *IEEE Trans. Power Electron.* 32(5), 3301–3316 (2017)
- [7] Wireless Power Consortium: The Qi Wireless Power Transfer System Power Class 0 Specification, Parts 1 and 2: Interface Definitions, Version 1.2.2. 1, 1–165 (2016)
- [8] Kao, J. J., Lin, C. L., Liu, Y. C., Huang, C. C., Jian, H. S.: Adaptive Bidirectional Inductive Power and Data Transmission System. *IEEE Trans. Power Electron.* 36(7), 7550–7563 (2021)
- [9] Wu, J., Zhao, C., Lin, Z., Du, J., Hu, Y., He, X.: Wireless Power and Data Transfer via a Common Inductive Link Using Frequency Division Multiplexing. *IEEE Trans. Ind. Electron.* 62(12), 7810–7820 (2015)
- [10] Yao, Y., Tang, C., Gao, S., Wang, Y., Alonso, J. M., Madawala, U. K., Xu, D.: Analysis and Design of a Simultaneous Wireless Power and Data Transfer System Featuring High Data Rate and Signal-to-Noise Ratio. *IEEE Trans. Ind. Electron.* 68(11), 10761–10771 (2021)
- [11] Ji, L., Wang, L., Liao, C., Li, S.: Simultaneous Wireless Power and Bidirectional Information Transmission With a Single-Coil, Dual-Resonant Structure. *IEEE Trans. Ind. Electron.* 66(5), 4013–4022 (2019)
- [12] Sun, Y., Yan, P. X., Wang, Z. H., Luan, Y. Y.: The Parallel Transmission of Power and Data With the Shared Channel for an Inductive Power Transfer System. *IEEE Trans. Power Electron.* 31(8), 5495–5502 (2016)
- [13] He, X. et al.: Wireless Power and Information Dual Transfer System via Magnetically Coupled Resonators. *Commun. Eng.* 3(1), 8 (2024)
- [14] Qian, Z., Yan, R., Wu, J., He, X.: Full-Duplex High-Speed Simultaneous Communication Technology for Wireless EV Charging. *IEEE Trans. Power Electron.* 34(10), 9369–9373 (2019)
- [15] Cheng, H., Zhang, L., Wang, L., Yao, Y., Mai, J., Lu, K., Xu, D.: A FDM-based Simultaneous Wireless Power and Data Transfer (SWPDT) System Functioning with High-rate Full-duplex Communication. *Proc. 22nd Int. Conf. Electr. Mach. Syst. (ICEMS)*, Harbin, China, 1–5 (2019)
- [16] Fan, Y., Sun, Y., Dai, X., Zuo, Z., You, A.: Simultaneous Wireless Power Transfer and Full-Duplex Communication With a Single Coupling Interface. *IEEE Trans. Power Electron.* 36(6), 6313–6322 (2021)
- [17] Trautmann, M., Sanftl, B., Weigel, R., Koelpin, A.: Simultaneous Inductive Power and Data Transmission System for Smart Applications. *IEEE Circuits Syst. Mag.* 19(3), 23–33 (2019)
- [18] Jing, Y., Feng, W., Qiao, K., Yang, L., Wang, S., Lu, L.: Simultaneous Wireless Power and Data Transfer System With Full-Duplex Mode Based on LCC/CLC Resonant Network. *IEEE Trans. Power Electron.* 38(4), 5549–5560 (2023)
- [19] Yang, L., Huang, J., Feng, B., Zhang, F., Zhang, Y., Li, X., Jian, J., Wang, Z., Tong, X.: Undersea Wireless Power and Data Transfer System With Shared Channel Powered by Marine Renewable Energy System. *IEEE J. Emerg. Sel. Topics Circuits Syst.* 12(1), 242–250 (2022)
- [20] Wang, P., Sun, Y., Feng, Y., Feng, T., Fan, Y., Li, X.: An Improvement of SNR for Simultaneously Wireless Power and Data Transfer System with Full-Duplex Communication Mode. *IEEE Trans. Power Electron.* (2021)
- [21] Xu, Q., Deng, J., Jiang, L., Fu, N., Wang, S.: A Half-Duplex Simultaneous Wireless Power and Data Transfer System and Its Design Method Based on Amplitude Shift Keying. *Proc. 6th Int. Conf. Energy Electr. Power Eng. (CEEPE)*, Guangzhou, China, 1482–1487 (2023)
- [22] Liu, Q., Xiao, J., Wu, X., Wu, N., Mo, Y., Xu, H.: An Improvement of SNR for Simultaneous Wireless Power and Data Transfer System with Full-duplex Mode Based on Dual-tuned Filter. *Proc. IEEE Int. Conf. Electr. Autom. Comput. Eng. (ICEACE)*, Changchun, China, 1–7 (2024)
- [23] Da, C., Wang, L., Li, F., Tao, C., Zhang, Y.: Analysis of Undersea Simultaneous Wireless Power and 1 Mb/s Data Rate Transfer System Based on DDQ Coil. *IEEE Trans. Power Electron.* 38(10), 11814–11825 (2023)
- [24] Luo, Y., Yang, Y., Hong, H., Dai, Z.: A Simultaneous Wireless Power and Data Transfer System With Full-Duplex Mode for Underwater Wireless Sensor Networks. *IEEE Sensors J.* 24(8), 12570–12583 (2024)
- [25] Li, T., Sun, Z., Wang, Y., Mai, J., Xu, D.: Undersea Simultaneous Wireless Power and Data Transfer System With Extended Communication Distance and High Rate. *IEEE Trans. Power Electron.* 39(3), 2917–2921 (2024)