Analysis and Design of 6.78 MHz Wireless Power Transfer System for Robot Arm

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SUMMARY This paper presents a design method of a two-hop wireless power transfer (WPT) system for installing on a robot arm. The class-E inverter and the class-D rectifier are used on the transmission and receiving sides, respectively, in the proposed WPT system. Analytical equations for the proposed WPT system are derived as functions of the geometrical and physical parameters of the coils, such as the outer diameter and height of the coils, winding-wire diameter, and number of turns. Using the analytical equations, we can optimize the WPT system to obtain the design values with the theoretically highest power-delivery efficiency under the size limitation of the robot arm. The circuit experiments are in quantitative agreement with the theoretical predictions obtained from the analysis, indicating the validity of the analysis and design method. The experimental prototype achieved 83.6 % power-delivery efficiency at 6.78 MHz operating frequency and 39.3 W output power.

key words: Wireless power transfer, multi hop, class-E inverter, coil design, optimization

1. Introduction

In recent years, wireless power transfer (WPT) systems have attracted much attention [1]–[13]. The WPT system researches are intended for various applications, such as electric vehicles [1], [2], medical devices [3], and robot arms [4]–[6]. The target application of this paper is a robot arm. When the WPT system is implemented in a robot arm, we can reduce the risk of failure due to disconnection by making the joints wireless. The WPT system has a multi-hop structure because wireless couplings are applied to each joint of the robot arm [4]–[6].

One of the challenges is to reduce the volume and weight of the WPT system. When the WPT system is heavy, it may affect the motion of the robot arm. The high-frequency operation is one of the solutions to this problem. The higher the operating frequency, the smaller are the passive elements such as coupling coils and resonant capacitances. The operating frequency of the WPT system for the robot arm presented in the previous papers is around 100 kHz [4]–[6].

New semiconductor materials such as gallium nitride (GaN) have made it possible to implement WPT systems at ISM band frequencies such as 6.78 MHz [7], [8]. However, switching losses are a significant problem at megahertz frequencies.

The class-E inverter, a type of DC-AC inverter, is often used in megahertz-band WPT systems. The class-E inverter can suppress the switching losses by the class-E switching [12]–[18] which achieves zero voltage switching (ZVS) and zero derivative switching (ZDS) simultaneously.

Another factor for the efficiency loss of the high-frequency WPT system is the power loss in the coupling coils. In the previous researches, the coils are firstly designed, and then the WPT system circuits are created using the designed coils [4]–[6]. The coupling coil must have sufficient self-inductance to transmit the desired power. However, coils with high self-inductance also have high equivalent series resistance (ESR), degrading the power delivery efficiency. Besides, the output voltage depends on the turn ratio of the transmission and receiving coils. Therefore, it is desirable to consider the coil design and the WPT system design simultaneously because the coupling coil has a significant impact on the system performance[19]. The importance of considering the coupling coil in a multi-hop WPT system increases with the increase in the coil number.

This paper, which is an extended version of [14], proposes a design method of a two-hop WPT system, which is assumed to be implemented on a robot arm. The class-E inverter is used on the transmission section in the proposed system, and the class-D rectifier is adopted on the reception one. To obtain the component values for satisfying the class-E ZVS/ZDS conditions, we derive the analytical expressions of the waveforms of the two-hop WPT system. In the analytical equations, self-inductance, coupling coefficient, and ESR of the coils are described as functions of the geometrical and physical parameters of the coils, such as outer coil diameter, coil length, winding-wire diameter, and number of turns. By combining the waveforms and the analytical expressions for the coils, we can derive the theoretical maximum power delivery efficiency and the parameter values for obtaining it under the size limitation.
of the robot arm. Experimental results were quantitatively in agreement with the analytical predictions, demonstrating the validity and effectiveness of the proposed design strategy. The laboratory measurements showed 83.6 % of the power delivery efficiency with 6.78 MHz operating frequency and 39.3 W output.

2. WPT System for Robot Arm

2.1 Overview of Robot Arm

In traditional robot arms, electric power is supplied to the built-in motor through wires. However, there is a risk of failure due to disconnection because the wires are twisted and worn at the joints. To solve this problem, a wireless connection of power transfer at the joints has been considered.

Figure 1 shows a schematic diagram of the robot arm we are developing. The arm length of the robot arm is 25 cm. The robot arm designed in this project is supposed to be manufactured with acrylic resin material using a 3D printer. Therefore, there is no metal material in the robot arm. It is necessary to transmit electric power to the upper arm body from the pedestal as the power source of the second and third-joint motors. Therefore, we develop a two-hop WPT system in this paper. As shown in Fig. 1(b), a solenoid coil is placed around the axis of rotation of the robot-arm joint as a power transmission/reception coil. With this structure, the distance and angle between the coils can be kept constant regardless of the motion of the robot arm. Therefore, the coupling coefficients between the transmitting and receiving coils are considered to be always stable.

The coupling coils are mounted on the bobbin as shown in Fig. 1(c). If a round copper wire of 1 mm in diameter is wound on the bobbin, the maximum self-inductance of the single-layer solenoid coil is 5.96 µH. In the magnetic-resonance WPT, the resonant frequency of the coupling coil and the resonant capacitance is approximately equal to the transmission frequency. When we consider the size of the resonant capacitance, a high frequency such as 6.78 MHz in

2.2 WPT System Construction

Figure 2(a) shows a circuit model of the proposed WPT system. The proposed system consists of three sections: power transmission section, relay section, and power receiving section. The switching loss is reduced despite a high frequency of 6.78 MHz by applying the class-E inverter [15]–[18] in the transmission section and a class-D rectifier [20] in the reception section. The operating waveforms of the proposed system are shown in Fig. 3, where \( \theta = \omega t = 2\pi f t \) is the angular displacement and \( f \) is the operating frequency. If the pre-post regulation method is adopted for the WPT system, the WPT system works at the optimal condition.
Regardless of load variations [21]. Therefore, it is important to achieve the optimal design at the rated load resistance [13].

The class-E inverter is composed of DC input voltage $V_I$, input inductance $L_C$, GaNFET $S$ as switching component, shunt capacitance $C_S$, and resonant filter $L_1$-$C_1$. The resonant inductance $L_1$ becomes the transmission coil at the first coupling. When the input inductance is sufficiently high, the input current $I_1$ can be regarded as a direct current, as shown in Fig. 3(a). The driver signal $v_g$ drives the GaNFET with transmission frequency $f$ and a 50 % duty ratio. During the OFF state of the GaNFET, a pulsed switching voltage $v_S$ occurs, as shown in Fig. 3(a). In the class-E inverter, the switch voltage at the moment when the switch turns on is zero (ZVS), and its derivative is also zero (ZDS), which is called the class-E ZVS/ZDS. The class-E inverter suppresses the switching loss and leads to high-efficiency operation at high frequencies because of the class-E ZVS/ZDS. When $L_1$ resonates with $C_1$, the current $i_1$ flowing through $L_1$ becomes a sinusoid, as shown in Fig. 3(a). Note that the class-E ZVS/ZDS does not always occur in this circuit configuration. The component values must be derived to achieve the class-E switching.

The relay section consists of two coils and a capacitance. $L_2$ and $L_3$, shown in Fig. 2, are the reception coil of the first coupling part and the transmission coil of the second coupling part, respectively, which are placed with the 25 cm distance on the arm. The coils are connected by the parallel copper wires, whose interval is 2.6 cm, as shown in Fig. 1(a). The parasitic inductance of the copper wire is modeled as $L_f$, as shown in Fig. 2. In this paper, the magnetic-resonance

WPT is adopted for suppressing the power consumption of the coupling part. The capacitance $C_2$ has a role of resonating with $L_2$, $L_3$, and $L_f$.

The reception section consists of a resonant circuit $L_4$-$C_4$ and a current-driven class-D rectifier, which also achieves high-efficiency operation at high frequencies. $L_4$ is the power receiving coil of the second coupling part. The voltage across the rectifying diodes is a square wave, as shown in Fig. 3(b). Through the $C_f$ - $R_L$ low-pass filter, the DC output voltage can be obtained.

### 3. Analytical Model

The objective of the design is to optimize the two-hop WPT system with the highest power-delivery efficiency. Analytical equations of the proposed system are derived for optimizing the WPT-system design.

#### 3.1 Coupling Coil Model

Figure 1 shows a WPT system for the robot arm to be designed in this paper. The coupling coils are designed to fit into the bobbin shown in Fig. 1(c). Each coil is designed differently under the limitation of the bobbin size.

Figure 4(a) shows an overview of a solenoid coil, where $h$ and $d$ are the height and diameter of the coil, $N_i$ and $N_j$ are the number of winding turns and layers, and $w$ and $c$ are the bare wire diameter and the thickness of the insulating coating, respectively. $w$ is usually determined by the standard such as Square (SQ) and American Wire Gauge (AWG), and $c$ is also standardized, for example, in Japanese Industrial Standards (JIS). No current flows through the insulating coating. The winding copper wire is wound at an interval $p$. In this paper, the coils are wound evenly, namely

$$p = \frac{h - N_i(w + 2c)}{N_i - 1}. \quad (1)$$

In the wireless coupling part, two coils $i$ and $j$, which have a length of the gap $l_g$, are coupled by a magnetic field, as shown in Fig. 4(b).

Three electrical parameters model the wireless coupling parts: self-inductance $L$, coupling coefficient $k$, and ESR $r_L$ [13]. These electrical parameters can be expressed as functions of the geometrical and physical parameters of the coils.
By using the Nagaoka coefficient $\xi$ [22],
\[
\xi = \frac{4}{3\pi \sqrt{1-q^2}} \left[ 1 - \frac{q^2}{q^2} K(q) - \frac{1 - 2q^2}{q^2} E(q) - q \right],
\]
the self-inductance of the solenoid coil is expressed by
\[
L = \frac{\xi \mu_0 \pi d^2 N_i^2}{2h},
\]
where $\mu_0 = 4\pi \times 10^{-7}$ N/A$^2$ is the vacuum permeability. Besides, $K(q)$ and $E(q)$ in Eq. (2) are the complete elliptic integrals of the first kind and second kind, respectively, and
\[
q = \frac{1}{\sqrt{\frac{2^2}{\pi^2} + 1}}.
\]
From Eq. (2) and Eq. (3), we can see that the larger the coil diameter and the larger the number of turns, the larger the self-inductance value becomes.

The coupling coefficient between the coupling coils $i$ and $j$ can be expressed as
\[
k_{ij} = \frac{M_{ij}}{\sqrt{L_i L_j}},
\]
where $M_{ij}$ is the mutual inductance between the coils $i$ and $j$. The mutual inductance can be calculated by the Neumann’s equation [19], namely
\[
M_{ij} = \sum_{m=0}^{N_i-1} \sum_{n=0}^{N_j-1} \frac{\mu_0}{8\pi} \int_0^{2\pi} \int_0^{2\pi} d_i d_j \cos(\phi_i - \phi_j) d\phi_i d\phi_j \sqrt{d_i^2 + d_j^2 - 2d_i d_j \cos(\phi_i - \phi_j) + 4l_{mnn}^2},
\]
where $l_{mnn}$ is the distance between the $m_{th}$-turned winding of coil $i$ and $n_{th}$-turned winding of coil $j$.

The ESR of the solenoid coil can be predicted by the Dowell’s equations [23], namely
\[
r_L = R_{dc} X \left[ \frac{\sin(2X) + \sin(2X)}{\cosh(2X) - \cos(2X)} \right] + \frac{2(N_i^2 - 1) \sinh(X) - \sin(X)}{3 \cosh(X) + \cos(X)},
\]
where $R_{dc}$ is the DC resistance of the winding wire, which can be expressed as
\[
R_{dc} = \frac{\rho l_w}{\pi (\frac{w}{2})^2} = \frac{4\rho N_i (d + 2N_i w)}{w},
\]
where $\rho = 1.724 \mu\Omega/m$ is the copper resistivity, and
\[
l_w = \pi N_i [d + 2N_i (w + 2c)],
\]
is the winding-wire length. Additionally, $X$ in Eq. (7) is expressed as
\[
X = \left( \frac{\pi}{4} \right)^{\frac{1}{2}} \frac{w}{\sqrt{\frac{\rho}{\pi \mu_0 d} \sqrt{\frac{w}{p}}} \sqrt{\frac{w}{p}}}
\]

We see from the above equations, three electrical parameters of the coupling parts can be expressed by the geometrical and physical parameters of the coupling coils. In other words, it is possible to optimize the WPT system with simultaneous considerations of the coil and WPT-circuit designs through the equivalent circuit model.

### 3.2 WPT-system model

For simplifying the analysis, the following assumptions are given.

1. The GaNFET and the rectifying diodes work as ideal switching components. The parasitic capacitance of the GaNFET is absorbed into the shunt capacitance $C_s$.
2. The class-E inverter satisfies the class-E ZVS/ZDS conditions in Eq. (23).
3. The dc-feed inductance $L_C$ is high enough so that the current through the inductance is a direct current.
4. The cutoff frequency of $C_f - R_L$ is sufficiently low compared to the transmission frequency. Therefore, the output voltage $V_o$ includes no ripple.
5. The loaded quality factor in the inverter, which is defined as
\[
Q = \frac{\omega L_{eq1}}{R_{inv}},
\]
is sufficiently high to generate a pure sinusoidal resonant current $i_1$, where $L_{eq1}$ and $R_{inv}$ are the equivalent inductance and resistance, as shown in Fig. 2(e).
6. The GaNFET is in the ON and OFF states for $0 \leq \theta < \pi$ and $\pi \leq \theta < 2\pi$, respectively.

### 3.3 Equivalent resistance and reactance of the reception and relay sections

Figure 2 shows the equivalent circuit for the analysis. Because the input reactance of the class-D rectifier is zero, the rectifier can be modeled by the input resistance $R_i$ as shown in Fig. 2(b) [24], where
\[
R_i = \frac{2R_L}{\pi^2}.
\]
The impedance $Z_2$, defined in Fig. 2(b), is modeled by the equivalent resistance and inductance, which are
\[
R_{eq2} = \frac{k_L^2 \omega^2 L_3 L_4 (r_L + R_i)}{(r_L + R_i)^2 + (\omega L_4 - \frac{1}{\omega C_s})^2},
\]
and
respectively. The ESR of $L_3$ is placed in series with $L_3$. Therefore, to suppress the power loss of $r_{L_3}$, $R_{eq2}$ must be maximized. From Eq. (13), we obtain

$$\omega = 1/\sqrt{L_4 C_4}. \quad (15)$$

In this case, we have

$$R_{eq2} = \frac{k_3^2 \omega^2 L_3 L_4}{r_{L_4} + R_i} = \frac{\pi^2 k_3^2 \omega^2 L^3 L_4}{\pi^2 r_{L_4} + 2 R_L}, \quad (16)$$

and

$$L_{eq2} = L_3. \quad (17)$$

By similar analytical process, the equivalent resistance $R_{eq1}$ and the inductance $L_{eq1}$ is expressed as

$$R_{eq1} = \frac{\pi^2 k_3^2 \omega^2 L^3 L_1 L_2}{r_{L_2} + r_{L_3} + R_{eq2}} = \frac{\pi^2 k_3^2 \omega^2 L^3 L_2}{(r_{L_2} + r_{L_3}) \cdot (\pi^2 r_{L_4} + 2 R_L)} \quad (18)$$

and

$$L_{eq1} = L_1. \quad (19)$$

respectively, under the condition of

$$\omega = 1/\sqrt{(L_2 + L_3 + L_4) C_2}. \quad (20)$$

Finally, we have the equivalent resistance of the class-E inverter as

$$R_{inv} = R_{eq1} + r_{L_1}. \quad (21)$$

Equations (13)-(21) show that the equivalent resistance increases as the self-inductance, operating frequency, and coupling coefficient increase. In other words, the equivalent resistance is determined by the coil parameters and the output resistance.

### 3.4 Voltage transfer function of the proposed WPT system

Referring to [15], we analyze the class-E inverter in the steady-state. The amplitude of the inverter resonant current is

$$I_1 = -\frac{4 \cos \phi_{inv}}{\pi R_{inv}} V_I. \quad (22)$$

where, $\phi_{inv}$ is the phase-shift between the driving signal and the output voltage of the class-E inverter, as shown Fig. 3(a), which can be expressed as

$$\phi_{inv} = \pi + \tan^{-1}\left(-\frac{2}{\pi}\right). \quad (23)$$

Equation (23) is a sufficient condition for satisfying the class-E ZVS/ZDS conditions [24]. The input current $I_I$ is

$$I_I = \frac{2 \cos \phi_{inv}}{\pi} I_1 = \frac{8 \cos^2 \phi_{inv}}{\pi^2 R_{inv}} V_I. \quad (24)$$

The amplitude of the current $i_2$, flowing through the relay section, can be expressed by Eq. (25), at the bottom of this page.

Similarly, the amplitude of the current $i_4$, flowing through the rectifier-section coil, can be obtained from Eq. (26), at the bottom of this page.

Because the current $i_4$ is the input current of the class-D rectifier, the output voltage of the WPT system is expressed as [24]

$$V_o = R_L I_O = \frac{R_L I_4}{\pi}. \quad (27)$$

From Eqs. (23), (26), and (27), the voltage transfer function of the WPT system with achieving the maximum power-delivery efficiency can be obtained by Eq. (28), at the bottom of this page.

Equation (28) shows that the output voltage can be uniquely determined from the inductor parameters and the design specifications of input voltage, operating frequency, and load resistance.
4. Power-loss Analysis

Figure 5 shows the equivalent circuit model for the power-loss analysis. By achieving the class-E ZVS/ZDS conditions, the switching loss can be ignored. Therefore, the ESRs of the coupling coils and the losses incurred in the switching components are considered as power-loss factors in this paper. It is assumed that the locations of power losses, such as ESRs, threshold voltage of the diodes, switch ON-resistance, are small enough not to affect the voltage and current waveforms obtained at the previous-section analysis.

The conduction loss of the GaNFET is expressed as

\[
P_S = \frac{r_S}{2\pi} \int_0^{2\pi} i_s^2 d\theta
\]

where \( r_S \) is the ESR of the GaNFET. The power loss at the input inductance is obtained from

\[
P_{L_C} = r_{L_C} I_f^2,
\]

where \( r_{L_C} \) is the ESR of \( L_C \). Besides, the power losses at the coupling-coil ESRs are

\[
P_{L_j} = \frac{r_{L_j} I_j^2}{2}, \quad j = 1, 2, 3, \text{and } 4
\]

The power losses of the diodes occur due to the forward voltage \( V_{th} \) and ON-resistance of the diodes \( r_D \). Therefore, we have

\[
P_{D_j} = \frac{1}{2\pi} \left[ \int_0^{2\pi} V_{th} i_D d\theta + \int_0^{2\pi} r_D i_D d\theta \right]
\]

\[
= I_o V_{th} + \frac{\pi r_D I_o^2}{4}, \quad j = 1, \text{and } 2
\]

Because of the page limitation, we omit to show the resulting equations of the power-loss factors.

From the above power-loss factors, theoretical power-delivery efficiency can be derived from

\[
\eta = \frac{P_o}{P_o + P_{D_1} + P_{D_2} + P_S + P_{L_C} + P_{L_1} + P_{L_2} + P_{L_3} + P_{L_4}}.
\]

5. System Optimization

5.1 Problem statement

As mentioned in Sect. 3 and 4, the output voltage and the power-delivery efficiency are uniquely determined by the given coil parameters. The challenge in the design is how to optimize the coil design for obtaining the highest power-delivery efficiency with the rated output voltage.

In this paper, the parameter set of the coupling coils is defined as \( x = (N_{11}, N_{12}, N_{22}, N_{23}, N_{33}, w_3, N_{14}, N_{14}, w_4) \). We pick a winding wire from the 47 types of the polyester enameled copper wire with the JISC-3215-0-1 standard. Therefore, the vector \( x \) is regarded as a set of the discrete parameters.

For performing the optimization, we propose the evaluation function as

\[
F(x) = \alpha \eta(x) + \beta e^{-\gamma(1 - \frac{V_{th}(x)}{V_{om}})^2},
\]

where \( \alpha \) and \( \beta \) are weight coefficients related to the power-delivery efficiency and the output voltage, respectively and \( \alpha + \beta = 1 \). Figure 6 shows \( e^{-\gamma(1 - x)^2} \) as the fixed \( \gamma \). The maximum value of this function is one and the degree of sharpness is determined by \( \gamma \). In this paper, \( \gamma = 1000 \) is adopted for optimization. Because the maximum value of the power-delivery efficiency \( \eta(x) \) is also one, the maximum value of the evaluation function becomes one.

The coils are constructed within the bobbin restrictions of \( h_j = 1.5 \text{ cm} \) and \( r_j = 2.6 \text{ cm} \). As a result, the optimization of the WPT system design can be described as

\[
\max_x F(x),
\]

Subject to \( N_{ij} w_j < 20 \text{ mm}, N_{ij} w_j < h_j = 150 \text{ mm}, J_j(x) < 10 \text{ A/mm}^2, \ (j = 1, 2, 3, 4) \)

where \( J_j = I_j / \pi (w/2)^2 \) is the maximum current density of the windings.

5.2 Particle Swarm Optimization

In this paper, we adopted the Particle Swarm Optimization (PSO) algorithm, which is one of the heuristic optimization
algorithms [25], for the system-design optimization. The heuristic optimization algorithm is suitable for searching the optimal discrete parameters because the PSO does not use gradient information. Because the number of adjustable parameters in the PSO for determining the solution-search performance is relatively small, searching for the best parameter set easily and quickly is possible.

Figure 7 shows a flowchart of the PSO algorithm. The particle \( \nu \) at \( \kappa \)th iteration has two sets of \( \lambda^p_{\nu,\kappa} \) and \( \lambda^g_{\nu,\kappa} \), which mean the personal best and the global best. The personal best is the best parameter set of the particle \( \nu \) itself in the iteration history, and the global best is the best parameter set in the swarm in the iteration history.

The position of a particle \( \nu \) at \( \kappa \)th iteration is represented by the parameter vector of \( x_{\nu,\kappa} \). Besides, each particle moves its position in the parameter spaces according to the velocity vector \( v_{\nu,\kappa} \).

Namely, the position vector is updated as

\[
x_{\nu,\kappa+1} = x_{\nu,\kappa} + v_{\nu,\kappa+1},
\]

where the velocity vector is expressed as

\[
v_{\nu,\kappa+1} = a_1 v_{\nu,\kappa} + a_2 u_2 (\lambda^p_{\nu,\kappa} - x_{\nu,\kappa}) + a_3 u_3 (\lambda^g_{\nu,\kappa} - x_{\nu,\kappa}).
\]

In Eq. (36), \( a_1 \) is the inertia weight, \( a_2 u_2 \) and \( a_3 u_3 \) are the acceleration coefficients, whose typical values are \( a_1 = 0.729 \) and \( a_2 = a_3 = 1.4955 \) [25]. Additionally, \( u_2 \) and \( u_3 \) are the random numbers following the uniform distribution in the range of \([0, 1]\). Therefore, the acceleration coefficients

\[
\begin{array}{|c|c|c|c|}
\hline
\text{Turns} & N_t & \text{Layers} & N_L & \text{Diameter of the bare wire} \omega & \text{Thickness of the insulating coating} c \\
\hline
L_1 & 7 & 1 & 2.0 \text{ mm} & 0.030 \text{ mm} \\
L_2 & 8 & 1 & 1.8 \text{ mm} & 0.029 \text{ mm} \\
L_3 & 7 & 1 & 1.8 \text{ mm} & 0.029 \text{ mm} \\
L_4 & 5 & 1 & 2.0 \text{ mm} & 0.030 \text{ mm} \\
\hline
\end{array}
\]
In this paper, we have designed and implemented a two-
hop wireless power transfer (WPT) system for installing on a robot arm. The class-E inverter and the class-D rectifier are adopted on the transmission and receiving sides, respectively, in the proposed WPT system. Analytical equations for the proposed WPT system are derived as functions of the geometrical and physical parameters of the coils. Using the analytical equations, we can optimize the WPT system to obtain the design values with the theoretically highest power-delivery efficiency under the size limitation of the coupling coil size limitation. The circuit experiments were in quantitative agreement with the theoretical predictions obtained from all the analyses, indicating the validity of the analysis and design method. The experimental prototype achieved 83.6% power-delivery efficiency at 6.78 MHz operating frequency and 39.3 W output power.

Equations (2)-(6) in this paper cannot be adopted directly when the robot arm is made of metal because the metal material affects the parameters of the coupling part. In this case, it is necessary to establish model equations of the coupling coils including the metal effect, which should be addressed in future research.

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References


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