Research Article

An Improved Current-Doubler Rectifier for the Marine Controlled Source Electromagnetic Transmitter

Hongxi Song , Yiming Zhang, and Yu Zhang

Faculty of Information Technology, Beijing University of Technology, Beijing 100124, China

Correspondence should be addressed to Hongxi Song; shxbill@163.com

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High power marine controlled source electromagnetic transmitters have gained interest with applications in marine geological survey and mineral resources exploration. The direct current to direct current (DC-DC) converter that is typically used in marine transmitters has some issues, as the insulated-gate bipolar transistor (IGBT) tube cannot achieve zero-voltage switching (ZVS). In particular, lagging-leg switching cannot easily achieve ZVS. The conversion efficiency of the heat converter requires improvement. This paper proposes an improved current-doubler rectifier for the marine controlled source electromagnetic transmitter (ICDR-MCSET). Resonant inductance is increased and a blocking capacitor is added to the converter (DC-DC) circuit, where the converter can achieve ZVS in a wide load range. This results in the effective decrease of the heating temperature and the improvement of transformation efficiency. Saber software simulation and a 20 KW electromagnetic transmitter are used to verify the results, which show that the method is feasible and effective.

1. Introduction

There has been a recent increase of marine mineral exploration investment, which is expected to gradually move from offshore research to deep-sea exploration. There is a need to develop an electromagnetic exploration system that can control the deep-sea mineral resources with independent intellectual property rights. There is some current research that provides technical support for the exploration and the development of deep-sea mineral resources [1, 2]. The marine controlled source electromagnetic method that is used in detecting submarine oil and gas reservoirs has been improved over the years. This method can identify high resistivity reservoirs, which will improve the drilling success rate. Current marine geophysical exploration technology services such as Electromagnetic Geoservices (EMGS) and WesternGeco or international oil companies such as Exxon Mobile, Statoil, or Shell have conducted nearly 1000 marine electromagnetic exploration projects [3]. The electromagnetic sounding system uses electromagnetic sounding transmitters towed by tugs to excite the electromagnetic waves at the bottom of the sea. A multicomponent electromagnetic receiver is placed at the bottom of the ocean to measure the electromagnetic

field values. The distribution characteristics of underground electricity are detected by calculating the apparent resistivity and phase, or by directly using the observed electric field and magnetic field. This allows the distribution pattern of marine bottom structure and mineral resources to be observed [4, 5]. Strong electromagnetic fields are formed at the bottom of the ocean using emission electrodes to emit high currents. The signal acquisition station then reclaims and analyzes these reflected electromagnetic signals so that the mineral resources, textures, and landforms of the seabed can be identified [6, 7].

Current marine electromagnetic detecting transmitters have many problems, including large volume, heavy mass, low efficiency, and poor transmitting waveforms, and they do not meet the current needs of actual exploration in the oil industry. The controlled source circuit is the core component of the marine electromagnetic sounding transmitter, which directly affects the performance and efficiency of the whole device [8, 9]. The coaxial winding transformer has been used, where the leakage inductance of the transformer eliminates the voltage spike on the secondary coil. This expands the ZVS range of the load, and an equivalent model of the current-doubler rectifier is obtained and analyzed [10]. A novel

FIGURE 1: The marine controlled source electromagnetic transmitter overall structure.

welding machine has been proposed [11], which uses a diode instead of the rectifier. This can effectively reduce the output ripple current. In [12], soft switching is proposed to build a small signal model ocean electromagnetic transmitter, which improves transmitter efficiency and reduces overheating. A resonant inductor and two clamping diodes are added in the phase shift full-bridge direct current to direct current (DC-DC) converter, greatly reducing the loss of insulated-gate bipolar transistor (IGBT) tube and high-frequency diode. A larger resonant inductor is used for the converter to achieve the ZVS at light load, but it easily causes the duty cycle to be lost; the analysis and experimental verification are given in this paper [13, 14]. In [15], the zero-voltage and zero-current switching (ZVZCS) full-wave rectification DC-DC circuit is proposed; it improves the efficiency of the traditional marine controlled source electromagnetic transmitter (T-MCSET) and reduces the duty cycle loss and expands the adaptive range of the load.

As is seen in Figure 1, the marine transmitter is divided into a ship portion and an underwater portion. The ship portion contains a ship deck generator, a rectifier circuit, a personal computer (PC) monitoring unit, and an optical transceiver. The underwater portion contains the underwater tow, the DC-DC controlled source circuit, the launching bridge, the underwater controller, and the transmitting electrode [12].

The shipboard generators provide the initial electrical energy for the entire electromagnetic detection launching system. The deck rectifier filter circuit primarily enables the three-phase alternating current (AC) voltage generated by the shipboard generator to be converted into direct current (DC) voltage; the single-phase AC voltage is obtained by inverting the DC voltage; the transformer T_2 boosts the AC voltage, and this provides high-voltage single-phase AC power for the underwater pressure cabin, so that the losses produced by the transmission of the cable from the ship to the underwater transmitter are reduced. The monitoring unit of the upper deck establishes remote communication by the tow cables and the underwater launch system control units when the cabin pressure is initialized. This allows the underwater cabin pressure to be controlled and monitored. The module contains a global positioning system (GPS) module that can be used for the entire transmitter system.

The underwater shipboard tow is used in the mechanical connection between the ship and the pressure cabin to

make the power and signal transmission be carried out simultaneously. The single-phase AC voltage is converted into DC voltage by the underwater portion rectifier filter circuit, so the controlled source DC voltage is then converted into a frequency adjustable square wave alternating current by the DC-DC controlled source circuit, and the electric energy is stimulated into the sea medium via a transmitting electrode. The control unit chips of the launching system implement the digital signal process (DSP) to gain control of the DC-DC controlled source circuit and the square wave alternating current frequency, as well as the monitoring of the transmitter temperature [16]. The use of optical fiber communication establishes the communication connection between the transmitter and the host computer. The DC-DC controlled source circuit is an integral part of the ocean transmitter, where its performance and efficiency directly affect the performance of the entire transmitter. In the following study, resonant inductance is increased and the blocking capacitor is added to the converter (DC-DC) circuit. This converter easily achieves the ZVS, resulting in a reduced heating temperature and an improved conversion efficiency. So, the marine transmitter works stably for a long time and produces a more stable waveform.

2. Marine Controlled Source Circuit

2.1. Traditional Marine Controlled Source Circuit. The traditional high power marine transmitter cannot fully achieve soft switching; the IGBT tube cannot achieve zero-voltage switching (ZVS); the converter efficiency is low, and it can overheat. The traditional phase shift full-bridge converter topology is seen in Figure 2.

2.2. Analysis of Hard Switching and Soft Switching Processes. Figures 3(a) and 3(b) show, respectively, hard switching turn-on process 1 and turn-off process 1, which belongs to fully hard switching; when the switch is turned on and turned off, the change of voltage and current is synchronous from the beginning to the end, resulting in mass energy consumption. Hard switching turn-on process 2 and turn-off process 2 are considered as semihard switching, where the voltage and the current changes do not begin simultaneously, resulting in less energy consumption (Figures 3(c) and 3(d)). The traditional transmitter switching working condition is considered semihard switching. Figures 3(e) and 3(f) depict

FIGURE 2: Traditional phase shift full-bridge converter topology.

FIGURE 3: Switching turn-on and turn-off processes.

soft switching turn-on process 3 and turn-off process 3. Before the switch is turned on, the voltage drops to zero, and the current drops to zero again before the switch is turned off.

Hard switching has the following major problems [17, 18]:

- (1) There are large switching losses. The switching of voltage and current waveform results in switching loss, which goes up rapidly with the increase of switching frequency.
- (2) When the switch is turned off, there is a large voltage spike. The switching frequency becomes higher, the switch is turned off faster, the induction voltage increases much more, and device breakdown is more likely to happen.
- (3) The current peak of the switch is large. The higher the frequency, the larger the current spike, resulting in overheating damage.

FIGURE 4: Several rectifier circuit diagrams.

- (4) Electromagnetic interference is serious. The higher the frequency, the larger the di/dt and dv/dt in the circuit, which leads to an increase in electromagnetic interference (EMI). This affects the normal operation of the rectifier and the surrounding electronic equipment.

These problems have limited the operating frequency of the switching device. Soft switching technology is an effective way to circumvent these problems. In this paper, an improved current-doubler rectifier is proposed for marine controlled electromagnetic transmitters. Resonant inductance is increased and a blocking capacitor is added to the converter (DC-DC) circuit in order to achieve the soft switching controlled source circuit. This converter can achieve ZVS in a wide load range while improving the efficiency of the converter. Overheating is reduced, and the control accuracy of the output voltage and current is enhanced. The controlled source circuit is then analyzed. There are 12 operating states within 1 cycle of the soft switching controlled source circuit [19, 20]. A simulation and a 20 KW transmitter are used to verify this method, which seems feasible and effective.

2.3. Circuit Comparison. There are several rectifier circuits as shown in Figure 4 [21].

Some common rectifier circuits are shown in Figure 4. The current-doubler rectifier circuit is shown in Figure 4(a). The full-wave rectifier circuit is shown in Figure 4(b). The full-bridge rectifier circuit is shown in Figure 4(c). The commonly used current-doubler rectifier circuit is compared with the full-wave rectifier circuit. The current-doubler rectifier circuit has the following advantages: the current effective value of transformer secondary winding is reduced, and the utilization ratio of the transformer is high; without center tap, the output current ripple is reduced to 1/2 of the secondary winding current ripple of the transformer, and the frequency of the current ripple is doubled. Therefore, the filter inductance and capacitance needed for the output voltage and current can be smaller, so the current-doubler rectifier circuit is also more suitable for distributed power dissipation requirements in the improved current-doubler rectifier for the marine controlled source electromagnetic transmitter (ICDR-MCSET), and it also has a good heat

dissipation function. The current-doubler rectifier circuit is compared with the full-bridge rectifier circuit; the number of diodes used in the current-doubler rectifier circuit is reduced to half. However, the current-doubler rectifier circuit uses more than one output filter inductor, and the structure is slightly complicated.

3. The Improved Circuit Structure and Working State Analysis

3.1. Improved Circuit Structure. As is shown in Figure 5, this section analyzes the 12 operating states of the phase shift full-bridge current-doubler rectifier. The analysis assumes the following conditions [17, 22]:

- (1) All switching tubes, diodes, inductors, capacitors, and transformers are ideal components.
- (2) $C_1 = C_2 = C_{12}$, $C_3 = C_4 = C_{34}$, and $L_1 = L_2 = L_0$.
- (3) The output filter capacitor C_0 is large enough.

3.2. The Improved Circuit Working State Analysis. The main waveforms of the DC-DC converter are seen in Figure 6, which help us analyze the circuit structure. It is assumed that the controlled source circuit satisfies the following conditions [10, 19, 20, 23–25].

(1) *Switching Mode 1*, $t_0 < t < t_1$. According to the equivalent circuit model (Figure 7(a)), during $t_0 < t < t_1$, S_1 and S_4 are turned on. The transformer primary current i_p flows through S_1 , the transformer primary winding, the blocking capacitors, C_b , and S_4 . The diode D_6 is turned on, and the diode D_5 is turned off. N is the transformer winding ratio, where at t_0 the currents of the filter inductances L_1 and L_2 are i_{L1} and i_{L2} , respectively. i_L is the sum of i_{L1} and i_{L2} . The expressions of the load current I_0 and the primary current i_p are as follows:

$$\begin{aligned} i_p(t) &= \frac{i_{L1}(t)}{N}, \\ i_L &= i_{L1} + i_{L2}, \\ I_0 &= i_L. \end{aligned} \quad (1)$$

FIGURE 5: The improved phase shift full-bridge converter circuit diagram.

FIGURE 6: Main waveforms of the DC-DC phase shift full-bridge converter.

(2) *Switching Mode 2*, $t_1 < t < t_2$. According to the equivalent circuit model (Figure 7(b)), at t_1 , S_1 is turned on, and the current i_p is charged to C_1 and is discharged to C_2 . The rising rate of the voltage U_{S1} is limited by the capacitances C_1 and C_2 , so S_1 is turned off under the ZVS mode. The i_p current is charged to the C_b blocking capacitors. During this period of time, $i_p = i_{L1}/N$. The voltages of the capacitors U_{C1} and U_{C2} are represented as follows:

$$\begin{aligned} U_{C1}(t) &= \frac{I_P(t_1)}{2C_{13}}(t - t_1), \\ U_{C2}(t) &= U_{in} - \frac{I_P(t_1)}{2C_{12}}(t - t_1). \end{aligned} \quad (2)$$

(3) *Switching Mode 3*, $t_2 < t < t_3$. According to the equivalent circuit model (Figure 7(c)), after D_2 and S_2 are turned on, the

$$\omega = \frac{1}{\sqrt{L_r C_b}}. \quad (3)$$

(4) *Switching Mode 4*, $t_3 < t < t_4$. According to the equivalent circuit model (Figure 7(d)), during this time, D_2 and S_4 continue to be turned on, as $U_{ab} = 0$, D_6 is turned off, and D_5 is turned on. The two-voltage inductance filter is $-U_o$, and the current of the two filter inductors is continuously decreased linearly when $i_o = -i_{L2}$, $i_p = -i_{L2}/N$.

(5) *Switching Mode 5*, $t_4 < t < t_5$. According to the equivalent circuit model (Figure 7(e)), at t_4 , S_4 is turned off, and the current i_p is charged to C_4 and is discharged to C_3 . The rising rate of the voltage U_{S4} is limited by the capacitances C_3 and C_4 , so S_4 is turned off under the ZVS mode. The current i_p is charged to the blocking capacitors C_b and the voltage U_{Cb} is raised. i_p can be considered as a constant current source, where the voltages of the capacitors C_3 and C_4 are represented as

$$U_{C4}(t) = \frac{I_p(t_4)}{2C_{34}}(t - t_4), \quad (4)$$

$$U_{C3}(t) = U_{in} - \frac{I_p(t_4)}{2C_{34}}(t - t_4).$$

(6) *Switching Mode 6*, $t_5 < t < t_6$. According to the equivalent circuit model (Figure 7(f)), when D_3 and S_3 are turned on, the dead time (t_d) between the S_3 and the S_4 drive signals should be larger than t_{45} . Although S_3 is turned on at this time, it does not flow through the current. The primary side current i_p is circulated by the diode D_3 . i_p is decreased linearly and the voltage U_{Cb} continues to rise.

(7) *Switching Mode 7*, $t_6 < t < t_7$. According to the equivalent circuit model (Figure 7(g)), S_2 and S_3 are turned on, i_{L1} is decreased, i_{L2} is increased, i_p is increased in the negative direction, and the voltage U_{Cb} continues to decrease. At t_7 , the converter begins the other half of the cycle $t_7 < t < t_{13}$, where the working condition is similar to the previous half cycle $t_1 < t < t_7$.

3.3. Soft Switching Implementation Conditions. Leading-leg switching is performed when the output filter inductance is at the maximum current (t_2 and t_8) stored energy. The lagging-leg switching is performed when the output filter inductance is at the minimum current (t_4 and t_{10}). The main waveform of the phase shift full-bridge converter demonstrates that the average value of the output filter inductor current is [10, 11]

$$i_{Lo} = \frac{i_{Lmax} + i_{Lmin}}{2}. \quad (5)$$

The maximum and minimum values of the output filter inductor current are i_{max} and i_{min} ; when $t_2 < t < t_{10}$, i_{L1} is dropped from i_{max} to i_{min} . This is explained as follows:

$$i_{Lmax} - i_{Lmin} = \frac{U_o}{L_o} \left(1 - \frac{D_e}{2}\right) T_0. \quad (6)$$

TABLE 1: Experimental circuit parameter design.

Parameters	Values
Input voltage (U_{in})	540 V
Output voltage (U_o)	100 V
Output current (I_o)	200 A
Switching frequency (f_o)	20 kHz
The ratio of the primary and secondary sides of the transformer (N)	5:1
Filter inductance ($L_1 = L_2$)	47 μ H
Load (R_o)	0.5 Ω
Filter capacitor (C_o)	1500 μ F
Absorption capacitance ($C_1 = C_2 = C_3 = C_4$)	0.47 μ F
Blocking capacitor (C_b)	20 μ F
Resonant inductance (L_r)	10 μ H

T_0 is the switching cycle, and D_e is the effective duty cycle of the converter.

$$i_{Lmax} = i_{Lo} + \frac{U_o(2 - D_e)T_0}{4L_o}, \quad (7)$$

$$i_{Lmin} = i_{Lo} - \frac{U_o(2 - D_e)T_0}{4L_o}.$$

Formula (7) demonstrates that i_{Lmin} is negative, where the heavier the load is, the larger the value of i_{Lmax} is and the smaller the absolute value of i_{Lmin} is. This means that the zero-voltage switches are easier to be achieved in heavy load than in light load for the leading-leg switching. The lagging-leg switching achieves zero-voltage switching under the heavy load, which is more difficult than under light load. The leading-leg switching utilizes the zero-voltage switching more efficiently than the lagging-leg switching.

4. Experimental Results

To verify the theoretical analysis of the ICDR-MCSET, the blocking capacitor is added and the resonant inductance is amplified in the circuit, so that simulation and physical experiments can be performed.

4.1. Simulation Verification. The 20 KW ICDR-MCSET is verified by the Saber simulation and the parameters are set as follows (Table 1).

The four IGBT tubes' drive wave is a phase shift full-bridge converter drive wave in the simulation (Figure 8). When the load current is 200 A (Figure 9), the lagging-leg switching voltages U_{S3} , U_{S4} , and U_{ab} do not show a voltage rise or fall caused by the shock peak. This demonstrates that the converter is soft switching and the simulation is effective. In Figure 10, when the current-doubler circuit is adopted, the output current ripple is reduced to 1/2 of the secondary winding current ripple of the transformer, and the frequency

FIGURE 8: S_1 , S_2 , S_3 , and S_4 IGBT tubes drive wave. FIGURE 9: U_{S3} represents the voltage between the collector and the emitter of the 3-IGBT tube. U_{S4} represents the voltage between the collector and the emitter of the 4-IGBT tube. The primary winding voltage is U_{ab} , the primary winding current is i_p , and the blocking capacitor is C_b .

of the current ripple is doubled. The output voltage and the current load ripple are small with high stability (Figure 11).

4.2. Physical Verification. The 20 KW ICDR-MCSET is verified by physical verification; Figure 11 is the physical picture of the ICDR-MCSET. The whole circuit is installed in a 70 cm

(diameter) * 190 cm (length) hermetic pressure cabin, with two ends connected with a single AC power supply and emission dipole. The IGBT module consists of two parts: module A and module B (Figure 12).

The four IGBT tubes' drive wave is a phase shift full-bridge converter drive wave in physical experiments

FIGURE 10: Two inductor currents are i_{L1} and i_{L2} , which are superimposed on current i_L .

FIGURE 11: Load current = I_0 and load voltage = U_0 .

FIGURE 12: Marine controlled source transmitter circuit physical diagram.

FIGURE 13: S_1 , S_2 , S_3 , and S_4 IGBT tubes drive wave.

FIGURE 14: (a) The waveforms of the primary and the secondary voltage of the transformer are not in the ZVS mode. (b) The waveforms of the primary and the secondary voltage of the transformer are in the ZVS mode.

(Figure 13). According to Figure 14(a), the experiment shows that, in the blue ellipse, the primary winding of the transformer is not achieved in the ZVS mode from the T-MCSET, and it belongs to the semihard switching state. When the resonant inductance is increased and the blocking capacitor is added, the freewheeling current in the circuit is relatively large, its duration is long, and the corresponding diode is turned on and continues to flow current. The IGBT tube voltage begins clamping when the voltage is zero, so the primary voltage U_{ab} does not exhibit a second buffer in the process of rising. As is shown in Figure 14(b), the primary winding of the transformer successfully achieves soft switching inside the blue ellipse from the ICDR-MCSET.

The lagging-leg switching U_{S4} is turned on (Figure 16) and does not appear during the rise of the second buffer. As is shown in Figures 14 and 16, the lagging-leg switching U_{S4} successfully achieves soft switching and so the primary winding voltage U_{ab} of the transformer does not appear during the rise of the second buffer. As is shown in Figures 15 and 16, the two inductance currents are almost symmetrical. The two diode currents achieve natural commutation. The two inductance currents and the two diode currents are slightly different; these may be due to the fact that the internal

parameters of the control circuit and the device are not completely consistent, which does not affect the experimental results.

The two-in-one IGBT modules (A and B) form a phase shift full-bridge converter. The temperature curve can be seen from Figure 17. When the transmitter continues operation for 5 hours, the temperatures of the transformer and IGBT modules are gradually increased with time. When the transmitter stays in operation for about 3.5 hours, the temperature of the T-MCSET transformer rises to 68°C as seen in Figure 17(a). However, the temperature of the ICDR-MCSET transformer rises to 52°C as seen in Figure 17(b), and all temperatures have a stable trend. The temperature of the transformer is in a controllable range. Module A is shown as a red curve and module B is shown as a blue curve in Figures 17(c) and 17(d). When the transmitter operates for about 1 hour, the temperature of the T-MCSET IGBT modules rises to 42°C as seen in Figure 17(c). However, the temperature of the ICDR-MCSET IGBT modules rises to 39°C as seen in Figure 17(d), and all temperatures have a stable trend. The temperature of the IGBT modules is in a controllable range.

According to the above analysis from the ICDR-MCSET, the primary voltage U_{ab} of the transformer does not exhibit a

FIGURE 15: i_{L1} is the inductance L_1 current, i_{L2} is the inductance L_2 current, and i_L is the current superposition of the inductance L_1 and the inductance L_2 .

FIGURE 16: Two rectifier diodes: i_{D5} and i_{D6} . The lagging-leg switching voltage: U_{S4} .

FIGURE 17: Temperature curve of the MCSET key components.

FIGURE 18: The transmitter emitting the voltage waveform V_{op} and the current waveform I_{op} .

FIGURE 19: The contrast curve diagram of the efficiency and the load current.

second buffer in the process of rising; the lagging-leg switching successfully achieves soft switching; and the temperature, conduction loss, heating capacity, and damage rate of the transformer and the IGBT modules are all reduced, extending their service life and improving the conversion efficiency of the controlled source circuit. The results show that the temperature of the ICDR-MCSET transformer and IGBT modules is lower than that of the T-MCSET.

Good device cooling and extending the service life of the transformer and IGBT modules are an important issue in the design of the ICDR-MCSET; the heat of the transformer and IGBT modules is too high, and this will lead to changes in these devices' operating curve; excessive temperatures can cause damage to these devices.

As is seen in Figure 18, the marine controlled source transmitter emits a current and a voltage waveform of 1 Hz at the sea floor. This waveform exhibits a strong stability, good linearity, and high controllability. This substantiates the feasibility of the ICDR-MCSET.

As the load current increases, the conversion efficiency of the ICDR-MCSET rises (Figure 19). The maximum efficiency is 93% better than the T-MCSET. The experimental results demonstrate that the temperature of the IGBT tube decreases,

the heat loss is reduced, and the efficiency of the DC-DC converter is improved in ZVS mode.

5. Conclusion

- (1) This paper proposes an improved current-doubler rectifier for the marine controlled source electromagnetic transmitter (ICDR-MCSET). The blocking capacitor is added and the resonant inductance is increased within the MCSET. The zero-voltage switching condition of the IGBT tube is analyzed, the expression of achieving the zero-voltage switching is given, and the corresponding formula is deduced. The appropriate saturated inductance is chosen to obtain high conversion efficiency. Finally, the lagging-leg switch U_{S4} successfully achieves soft switching from the waveforms of the Saber simulation and the ICDR-MCSET. This shows that this method is feasible and effective.
- (2) Under the same condition, the temperature recorder is effectively used to record the temperatures of the transformer and IGBT modules in the laboratory. The results show that the temperature of the ICDR-MCSET transformer and IGBT modules is lower than that of the T-MCSET. The efficiency of the ICDR-MCSET is improved by about 1–3% under the entire load condition when compared to the T-MCSET, so conduction loss, heating capacity, and damage rate of the transformer and IGBT modules are all reduced, extending their service life, and the efficiency of the ICDR-MCSET is further improved.
- (3) Due to the small and hermetic space of the underwater pressure cabin, the emission power of the ICDR-MCSET is relatively large and is 20 KW; besides, the ICDR-MCSET needs to run for a long time; the heat dissipation and layout requirements of the devices are relatively high in the underwater pressure cabin to improve the power density of the ICDR-MCSET.
- (4) The Saber simulation and the ICDR-MCSET are used in the laboratory to verify the conclusion. The experimental waveform and analysis are given, and the developed marine electromagnetic detection system draws on the latest technology of switching power supply, in order to obtain high stability, high linearity, high power density, and high transmission efficiency of the ICDR-MCSET. The stable emission current will take marine electromagnetic survey to a new level.

Conflicts of Interest

The authors confirm that this paper's content has no conflicts of interest.

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