Excitation circuit for fluxgate sensor using saturable inductor

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Abstract

Traditional excitation circuits for fluxgate sensors consist of a lossy resistor or a bulky inductor (operated in non-saturated mode), whose function is to limit the excitation current. As a result, a high voltage source is required to provide sufficient excitation current to the sensor. For these reasons, traditional excitation circuits are unsuitable for portable battery operated devices that use fluxgate sensors. In order to reduce the circuit size and operating voltage in such applications, the use of a saturable inductor is proposed in this paper. Practical methods for choosing suitable circuit components are described to ensure that the circuit operates under optimum conditions. Experimental results illustrate a good linear relationship between the second harmonic voltage and the measured current for a fluxgate dc current sensor using the proposed excitation circuit.

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1. Introduction

Fluxgate sensors are widely used in magnetic field and current measurements. The excitation current source for a fluxgate sensor typically consists of an ac voltage source connected in series with a large output impedance. The impedance must be larger than the impedance of the fluxgate sensor so that the excitation current is independent of the impedance of the sensor. It is usually realized with either a lossy resistor [1] or a bulky inductor [2] operated in non-saturated mode. When the excitation frequency decreases, the inductor size increases to achieve a sufficiently high impedance. Since the current delivered from the excitation circuit is inversely proportional to the output impedance value, a high voltage source is needed to supply sufficient current to saturate the fluxgate sensor. Clearly, traditional excitation methods present a barrier to miniaturization of portable equipment that use fluxgate magnetic sensors, such as electronic compasses and ammeters.

This paper proposes a method for reducing the size and operating voltage of fluxgate excitation circuits by replacing the large non-saturated inductor with a saturable inductor.

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2. Characteristics of the fluxgate current sensor and selection of resonant capacitor C r

A small ferrite ring core can be adopted because the inductor is allowed to operate in the saturated region. The optimal operating conditions of the circuit are investigated and described. Circuit analysis is simplified by approximating a nonlinear inductance with an effective inductance value. Practical means of component selection for maximizing circuit stability and linearity are detailed. The sensitivity of an electroplated fluxgate current sensor using the proposed excitation circuit is measured to evaluate the circuit performance.

The proposed excitation circuit is demonstrated for a planar fluxgate current sensor. Two layers of toroidal magnetic core, each with thickness of 20 μm, are electroplated and embedded into a PCB. The outer diameter and the width of the core are 10 and 1 mm, respectively. The excitation coil has 40 turns. The coil tracks are etched on the PCB and copper through-holes complete the turns around the toroidal core. The excitation coil is connected to an ac excitation current source, where the amplitude of the current is high enough to deeply saturate the magnetic core. A hole is drilled in the PCB at the center of the toroidal core, through which a wire carrying the measurement current is passed. The mea-
From the denominator of (2), the damping factor, $\xi$, or

$$\xi = \frac{R_s}{\sqrt{L_sC_r}} \text{ or } \frac{R_s}{\sqrt{L_C}}$$

From (3), the damped resonant frequency is

$$f_r = \frac{1}{2\pi} \sqrt{\frac{1}{L_sC_r} - \frac{R_s^2}{4L_s^2}}$$

When the fluxgate current is increasing and the magnetic core reaches the saturated region, its inductance value drops rapidly and causes the resonant capacitor, $C_r$, to discharge. To maximize the resonant peak current of the sensor under saturated conditions, the damping factor, $\xi$, should be smaller than unity. Therefore, from (4),

$$\frac{R_s}{\sqrt{L_sC_r}} < 1 \Rightarrow C_r < \frac{4L_s}{R_s^2}$$

where $L_s$ denotes the saturated inductance value of the fluxgate sensor.

The relationship between the core magnetic field and the applied magnetomotive force (mmf) is investigated in order to determine saturation conditions for the fluxgate core. The excitation coil of the fluxgate sensor is connected to the current source shown in Fig. 1. The frequency of the current source is 50 kHz. A temporary secondary coil of 20 turns is manually wound around the core for magnetic field measurement. The voltage induced on the secondary coil is measured with a digital oscilloscope, and the magnetic field inside the core is found by numerically integrating the secondary voltage instead of using a traditional analog integrator circuit. In this way, errors coming from the phase shift and tolerance of electronic components in the analog integrator circuit are eliminated. The measured core magnetic field versus applied mmf to the fluxgate sensor is shown in Fig. 2.

When the applied mmf to the sensor is higher than 8 A turn, (0.2 A in 40 turns), the magnetic core begins to saturate. There is no significant increase in magnetic field when the applied mmf is higher than $\sim 18$ A turn. In the proposed excitation circuit, the peak mmf applied to the fluxgate sensor is 21 A turn (0.525 A peak sensor current) to guarantee that the magnetic core is deeply saturated so that the measurements are repeatable [3].

Fig. 3 shows the inductance and resistance versus biased dc mmf of the magnetic sensor measured at 50 kHz. As in Fig. 2, it can be seen that when the applied mmf is approximately 8 A turn, the magnetic core inside the sensor begins to saturate. In order to ensure that the fluxgate operates in the deeply saturated region, a magnetomotive force of at least 21 A turn is applied to the sensor in this work. At this level, the inductance and series resistance (including the core loss) of the sensor are 0.86 $\mu$H and 1.5 $\Omega$, respectively. Substituting these values into (6):

$$C_r < \frac{1.5 \times 6}{4(0.86 \times 10^{-6})} \approx 1.5 \mu F$$

Therefore, a capacitance value of less than 1.5 $\mu F$ is desired to achieve a resonant peak sensor current.
The measurements in Fig. 3 are used to verify the calculation of $C_r$ above. By substituting the values of $L_s$ and $R_s$ from Fig. 3 into (5), with different resonant capacitance values (from 0.1 to 2 μF), the resonant frequency corresponding to the sensor operated under dc bias may be obtained. Results are plotted in Fig. 4. When the sensor current is 0.525 A, i.e. biased mmf is 21 A turn, the resonant frequency decreases with increasing resonant capacitance value. As predicted with (7), the results show that the resonant circuit is over-damped when the resonant capacitance is higher than 1.5 μF for this bias value.

On the lower limit, the capacitance value should be large enough to store energy and provide sufficiently high peak current to the sensor. The final choice of $C_r$ is determined by examining the stability of the transient response of the circuit for different values of resonant capacitance. Measurements of the sensor current waveform are presented for capacitance values varying from 0 to 2 μF in Fig. 5. The excitation current applied to the resonant network is a 50 kHz square current. The results show that a capacitance value larger than 0.33 μF is required to store sufficient energy. A peak sensor current of ~0.5 A can be achieved by using resonant capacitance values between 0.33 and 1.33 μF. As predicted above, the sensor peak current decreases significantly when the capacitance value is higher than 1.5 μF.

It was shown in Fig. 4 that the resonant frequency increases with decreasing capacitance value for any particular
Resonant frequency versus Biased mmf of the Current Sensor with Different Capacitance Values

![Graph showing resonant frequency vs. biased mmf for different capacitance values.]

Fig. 4. Resonant frequency vs. biased mmf of the fluxgate current sensor with different resonant capacitance values.

Fig. 5. Excitation current waveform of the fluxgate current sensor with different resonant capacitance values.

Biased mmf. Waveforms in Fig. 5 reveal that a high resonant frequency compared to the operating frequency causes unstable high-frequency oscillations. As shown, a capacitance value of $1 \mu F$ is the optimized value as it can achieve a high peak sensor current and does not cause undesired high-frequency oscillations.

### 3. Characteristics of the saturable inductor

The size of a non-saturated inductor required in a traditional current source circuit may be excessive for some portable electronic equipment. In this section, it is shown that it is not necessary for the inductor to operate in the...
non-saturated mode over the whole excitation cycle, so that a saturable inductor can be adopted to replace a more bulky non-saturated component.

The main function of the inductor is to limit the current drawn from the source for a given fluxgate excitation current. This can be achieved by driving the circuit so that the inductor (with larger impedance than the fluxgate sensor) operates in the non-saturated mode over part of the excitation period. For this purpose, it is important to ensure that the inductor is either non-saturated or saturated with current flowing in the previous direction when the polarity of source voltage is changing. In the non-saturated state, the high impedance of the inductor limits the current flowing from the source as in the case of a non-saturable component. In the case that the polarity of the square voltage is changing (e.g. from negative to positive) and the inductor is saturated with the previous current direction (i.e. negative direction), the current reduces to 0 A in a short time. The inductor therefore becomes non-saturated with impedance high enough to restrict the source current over the start of the cycle. Measured waveforms are presented later to illustrate this mode of operation. If switching occurs when the inductor is saturated with current flowing in the same direction as the voltage change (e.g. due to resonance), the inductor is driven more deeply into saturation after switching and its impedance is too low to limit the source current.

By allowing the inductor to operate in the saturated mode over part of the cycle as described, its size and weight can be reduced. In the prototype circuit, a small size ferrite ring core with high permeability ($\mu_r = 11,000$) is used. The outer diameter, width and height of the core are 1 cm, 2 and 4 mm, respectively. The number of turns is chosen as a trade-off between inductor impedance and (source) voltage level required to drive the inductor into saturation, as given by the induced emf, $v(t)$, across an inductor:

$$v(t) = \frac{d}{dt}B(t)$$

where $B(t)$ is the magnetic flux density in the inductor magnetic core, $N$ the number of turns of inductor winding, and $A$ is the effective cross sectional area of the magnetic core.

In frequency domain,

$$V(s) = sNAB(s)$$

(9)

From (9), the operating voltage of the saturable inductor is directly proportional to operating frequency, number of turns, effective area of the magnetic core and the saturation level of the magnetic flux density. When the material and dimensions of the magnetic core are fixed, both the operating frequency and number of turns of the inductor determine the operating voltage of the saturable inductor and hence that of the sensor excitation circuit. In the prototype excitation circuit operating at 50 kHz, four winding turns are used so that the excitation circuit can be operated at voltage levels as low as 3 V.

The measured core magnetic flux plotted against the applied mmf of the saturable inductor is shown in Fig. 6. Measurements are taken with the test circuit shown in Fig. 1, with the peak saturable inductor current set to 0.25 A. This is the peak current required to drive the fluxgate in the resonant circuit into deep saturation, as described in Section 2.

From Fig. 6, it is seen that the inductor begins to saturate when the mmf applied to the inductor increases to about 0.5 A. Unlike the fluxgate sensor, the saturable inductor need not be deeply saturated in this case. The level of saturation depends on the inductor current and the size of the magnetic core. According to the results in Fig. 6, the magnetic core is not deeply saturated so a magnetic core with smaller cross sectional area could be used. It is envisaged that the
magnetic core for the saturable inductor could be an electroplated planar core, embedded into a PCB and fabricated together with the electroplated fluxgate sensor.

4. Selection of the blocking capacitor \( C_b \)

The input impedance, \( Z_{in}(s) \), of the whole resonant circuit, as shown in Fig. 9, is given by

\[
Z_{in}(s) = \frac{1}{sC_b} + R_{sat} + sL_{sat} + Z_t(s)
\]

(10)

where \( C_b \) is the dc blocking capacitor, \( R_{sat} \) and \( L_{sat} \) are the series resistance and inductor of the saturable inductor, respectively, and \( Z_t(s) \) is the impedance of the fluxgate sensor in parallel with the resonant capacitor \( C_r \). Clearly, in addition to providing a dc blocking function, \( C_b \) contributes to the impedance of the circuit in this case. For efficient operation of a half-bridge converter circuit, an inductive load with dead-time or a soft-switching scheme is desired [5], so the choice of \( C_b \) is determined by the impedance of other circuit components at the operating frequency of the circuit.

Because the inductance values of the saturable inductor and the fluxgate sensor vary with the applied current, effective inductance values for the two inductive devices are necessary in order to simplify the analysis in (10). From [4], the effective inductance \( L_{eff} \) of a time-varying inductance \( L(t) \) can be expressed as

\[
\frac{1}{L_{eff}} = \frac{1}{T} \int_{0}^{T} \frac{1}{L(t)} dt
\]

(11)

where \( L(t) = v(t) - i(t)R/di(t)/dt \), \( v(t) \) and \( i(t) \) are the instantaneous voltage and current of the inductor, \( R \) the series resistance of the inductor, and \( T \) is the period of the ac excitation. The discrete form of the inductor voltage \( v_k \) and current \( i_k \) can be obtained by using a digital storage oscilloscope and the effective inductance can be expressed as

\[
\frac{1}{L_{eff}} = \frac{1}{N} \sum_{k=1}^{N} \frac{(i_{k+1} - i_{k-1})/\Delta t}{v_k - i_k R}
\]

(12)

where \( N \) is number of samples in one excitation period taken by the oscilloscope and \( 1/\Delta t \) is the sample rate of the oscilloscope.

When the fluxgate sensor is excited by the current source shown in Fig. 1 and with \( C_b \) of 1 \( \mu \)F, the instantaneous value, \( L_e(t) \), of the fluxgate sensor is measured. Its effective inductance value, \( L_{eff}(t) \), is calculated using (12) and equal to about 4.3 \( \mu \)H. The effective inductance, \( L_{eff}(t) \), of the saturable inductor is found to be 26 \( \mu \)H by measuring the device under the same conditions as described in Fig. 6. It should be noted that these excitation conditions are close to those applied in the actual sensor excitation circuit. By substituting the effective inductance values, optimized resonant capacitance value of \( C_r \) (1 \( \mu \)F) and different blocking capacitance values, \( C_b \) (from 0.2 to 1 \( \mu \)F) into (10), the magnitude and phase of the input impedance, \( Z_{in} \), of the resonant circuit are plotted and shown in Figs. 7 and 8, respectively. As shown, the resonant frequency of \( Z_{in} \) decreases as the value of \( C_b \) increases.

The dc supply voltage of the excitation is determined by the number of turns of the saturable inductor. When the number of turns has been determined, the input impedance magnitude \( Z_{in} \) and hence the sensor excitation current can be fine-tuned by the value of \( C_b \). It should be noted that the analysis using effective inductance is to approximate the nonlinear resonant circuit (due to non-linear inductors) to a linear LRC circuit. This method gives useful information
to predict the phase and resonant frequency of the circuit, but the exact excitation current should be verified experimentally. In the prototype circuit, a 3 V dc supply voltage is chosen and the peak current of the fluxgate sensor is set to about 0.525 A by selecting $C_b = 0.47 \mu F$. The resonant frequency is about 44 kHz so that the resonant circuit can be sufficiently inductive at a switching frequency of 50 kHz.

5. Operating principle of the excitation circuit

The circuit schematic of the proposed excitation circuit for a fluxgate sensor is shown in Fig. 9. It is similar to a commonly used fluxgate excitation circuit except that the non-saturated inductor is replaced by a small saturable inductor and the square voltage source is achieved by a simple half-bridge instead of a full-bridge converter followed by a step-up transformer [2]. As discussed above, the capacitor, $C_b$, blocks the dc component of the square voltage and also takes part in the resonant network. The fluxgate sensor is connected in parallel to the resonant capacitor, $C_r$, which provides a high peak current to the fluxgate [3].

The operating frequency of the circuit is set to 50 kHz, which is limited by the maximum frequency of the SR830 lock-in amplifier [6]. The dc supply voltage, $V_{DD}$, can be as low as 3 V, so low voltage rating (30 V) and low turn-on resistance (0.05 $\Omega$) IR7303 MOSFETs can be used in the half-bridge circuit. The IR7303 MOSFET module consists of two same type n-channel MOSFETs, which are fabricated in an SO8 surface mount package, so that their characteristics and operating condition, such as die temperature, can be almost identical. This arrangement has an advantage of increasing the symmetry of the output square voltage and so the undesired second harmonic content of the square voltage can be suppressed. The MOSFETs, $S_1$ and $S_2$, operate under zero voltage conditions so that energy dissipation due to conduction loss is minimized. The capacitance values of $C_b$ and $C_r$ are 0.47 and 1 $\mu F$, respectively. The measured dc supply current is about 63 mA and the power consumption of the excitation circuit, including the half-bridge and the resonant circuit, is about 190 mW.

The resonant frequency of the resonant circuit is designed to be lower than the operating frequency so that the input current, $I_{in}$, flowing into the resonant circuit does not complete half of the resonant period when the next half cycle begins. It implies that when $S_2$ is switched off and $S_1$ is switched on, the saturable inductor is saturated by the reverse current. This arrangement has the advantage that the current delivered from the half-bridge converter has a higher rms value for the same peak current value (or has lower peak current value for the same rms value).
When $S_2$ is switched off and $S_1$ is switched on, the output voltage of the half-bridge rises from 0 to 3 V. At the switching transition, the saturable inductor is saturated by the negative current. Its inductance value is very low so the current increases rapidly from negative to zero ampere after about 800 ns as shown in Fig. 10. In the time interval between $t_1$ and $t_2$, the inductor current becomes positive and is in the non-saturated region so the output current from the half-bridge converter can be limited. The inductor becomes saturated by the positive current at $t_2$, and at $t_3$ the current increases to its peak value due to the resonant effect. While the inductor current is still high and the saturable inductor is being saturated by the positive current, the negative half-cycle begins (i.e. $S_1$ is off and $S_2$ is on). Thus, a high input rms current value of the resonant circuit can be achieved with the same peak current value. The operation of the positive and negative half cycles is similar and the measured waveforms shown in Fig. 10 are symmetrical.

It should be noted that the saturable inductor should have a high non-saturated inductance value to restrain the abrupt increase of current flowing from the half-bridge converter to the resonant network. High current flowing through the MOSFETs could increase the voltage drop across the drain-to-source terminals of the MOSFETs and distort the output.
voltage of the half-bridge converter. Furthermore, a high rate of change of current could lead to noise and EMI problems that degrade the accuracy of the current sensor.

The measured voltage and current of the fluxgate current sensor are shown in Fig. 11. When \( S_1 \) is switched on, \( C_r \) is charged smoothly as the charging current is limited by \( L_{sat} \). At time \( t_4 \), the sensor current increases to about 0.2 A, the magnetic core inside the fluxgate sensor becomes saturated. The impedance of the fluxgate sensor then drops rapidly, the resonant capacitor \( C_r \) is discharged and desired high peak current of 0.525 A flows into the fluxgate sensor at \( t_5 \).

6. Sensitivity of the fluxgate sensor using the proposed excitation circuit

The response of the fluxgate operated as a current sensor, with a dc current flowing in a wire passed through the centre of the fluxgate core, is investigated in this section. The second harmonic voltage measured across the excitation winding is measured (twice) and plotted against test dc current in Fig. 12. The test current increases from 0 to 10 A and then back to 0 A, with 0.5 A current step. Afterwards, the polarity of the measurement current reverses, decreases to \(-10\) A and back to 0 A again. Since the excitation current deeply saturates the fluxgate magnetic core \([3]\), the measurements in Fig. 12 are repeatable. As the excitation current is symmetric, the measured second harmonic voltage has a negligible offset and is odd-symmetric about the origin. From Fig. 12, the linear region is between about 0 and \( \pm 1\) A. Measurements of the sensor response with a finer current step of 0.01 A are plotted in Fig. 13. In this case, the test current is from \(-2\) to 2 A. Results in both Figs. 12 and 13 show that the sensitivity of the sensor is 90 mV/A in the linear region.
7. Conclusions

The use of a saturable inductor in a current excitation circuit for a fluxgate sensor has been successfully demonstrated. The size, supply voltage and power consumption of the excitation circuit has been reduced significantly. These advantages are desirable in portable battery operated electronic equipment using magnetic sensors that are based on the fluxgate principle, such as a digital compass.

The optimum operating mode of the excitation circuit using the saturable inductor has been illustrated. Practical methods of choosing suitable blocking and resonant capacitors are devised to ensure that the excitation circuit operates in a stable and efficient manner. The excitation current is highly symmetrical and deeply saturates the magnetic core of the fluxgate sensor, so that the offset error of the second harmonic voltage is very small. In the circuit prototype, the saturable inductor is made of a ferrite ring core of 1 cm diameter with 4 turns winding. It is envisaged that the saturable inductor could be electroplated and integrated into a PCB together with the fluxgate sensor.

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References


Biographies

S.C. Tang was born in Hong Kong in 1972. He received the BEng degree (with first class honors) and the PhD degree in Electronic Engineering at City University of Hong Kong in 1997 and 2000, respectively. His research interests involve high frequency electromagnetism, low-profile power converter design and analog electronics. After he graduated, he worked as a Research Fellow at the same university and then joined the National University of Ireland Galway as a Visiting Academic in 2001. He is presently a Visiting Scientist in Massachusetts Institute of Technology. He has won several awards, the most recent being the champion of the Institution of Electrical Engineers (IEE), Hong Kong Younger Member Section Paper Contest 2008.

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