Detection of weak magnetic fields propagated through a ferrous steel boundary using a super narrowband digital filter

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Abstract. An instrumentation system is described that is capable of launching high frequency magnetic fields through a large mild steel plate 2 mm in thickness, and detecting them on the face opposite to the transmitter with remarkable signal to noise ratios. Results for signal frequencies ranging between 4.5 kHz and 13 kHz are reported. The skin depth at 9 kHz, for the steel used, is approximately 137 µm. The detection of the minute fields arriving at the receiving coil is made possible by the use of digitally synthesized input signals, low-noise amplification, and in particular the use of a real time digital signal processing system that isolates the signal of interest using a super-narrowband IIR filter and very high levels of distortion-free gain. Although traditional methods of weak signal detection, such as lock-in amplification, may also be applied in this context, the digital approach discussed here is both more cost effective and flexible, allowing the simultaneous detection of multiple frequencies.

1. Introduction

There are numerous occasions when it is desirable to transmit a time-varying magnetic field through a metal structure and detect the signal that has travelled through the material, either in reflectance or in through-transmission mode. Examples include nondestructive testing, remote field eddy current (RFEC) inspection, materials characterisation and security screening [1 - 5]. A time varying magnetic field will induce eddy currents in a target, such as a steel plate, which in turn generate magnetic fields throughout the plate thickness.

Due to the skin-effect however, which is defined as the tendency of a high frequency electric current to distribute itself inside a conductor so that the current density is greatest at the surface, the strength of both the eddy currents and the associated magnetic field fall rapidly with depth in ferrous materials. The equation which describes the fall in current density is given by

\[ J = J_s e^{-\frac{d}{\delta}} e^{i\omega \left(\frac{d}{\delta}\right)} \]  

(1)

Where \( J_s \) is the current density on the surface, \( d \) is the depth within the material, \( \delta \) is the skin depth, \( \omega \) is the angular frequency and \( J \) is the is the current density at depth \( d \). The skin depth for a given material is governed by the relationship
\[
\delta = \sqrt{\frac{2}{\omega \mu_0 \mu_r \sigma}}
\]

where \(\sigma\) is the conductivity of the conductor or target, \(\mu_r\) its relative permeability and \(\mu_0\) is the absolute permeability of a vacuum. The skin effect severely constrains the operational frequency range of eddy current equipment, a factor which limits its ability to detect and resolve small features at depth. In this paper however, we describe an instrumentation system that is capable of launching alternating magnetic fields through a mild steel plate 2 mm in thickness and detecting them on the face opposite to the transmitter with exceptional purity. Because material properties vary in practice, it is difficult to know the precise values of the conductivity and permeability of a mild steel sample and so obtain an accurate measure of its skin depth. Moreover, the permeability of magnetic materials is not constant, but changes over a several-decade range as the excitation level is varied. Often, the permeability value given in tables is the maximum permeability or the maximum relative permeability for a specific material. The maximum permeability is the point where the slope of the B/H curve for unmagnetised material is the greatest. This is often taken as the point where a straight line from the origin is tangential to the B/H curve. Here, the permeability has been chosen to be of a low value because, at the small applied field strengths, it is an incremental permeability associated with a minor hysteresis loop, rather than the typically tabulated maximum permeability of a full B/H loop.

Assuming therefore conservative figures for the conductivity as \(6.0 \times 10^6\ \text{Sm}^{-1}\) and a relative permeability of 250, a skin depth of 0.194 mm is obtained at 4.5 kHz and 0.137 mm at 9 kHz. From equation (1), it is clear that for an excitation frequency of 4.5 kHz the (eddy) current density at a depth of 2 mm is approximately equal to \(0.33 \times 10^{-4}\) of its value at the surface. Since the magnetic field is proportional to the current, the field radiated on the back face of the plate will experience similar levels of attenuation. For an excitation frequency of 9 kHz, the attenuation at a depth of 2 mm is approximately \(0.457 \times 10^{-6}\).

The success of the detection systems is predicated on the design and use of digitally synthesised, stable excitation signals, low-noise instrumentation amplifiers, a powerful real time digital signal processing (DSP) system that performs super-narrowband filtering, and a very high resolution analogue-to-digital and digital-to-analogue (codec) converter.

2. Instrumentation and experiment

Figure 1 depicts the instrumentation used to transmit and detect the high frequency magnetic fields. A 50 mV peak-to-peak sinusoidal signal at a single frequency was synthesized by a Thurlby Thandar TGA1230 digital function generator. This was fed to a current power amplifier, with a voltage gain of 30 dB, whose output was in turn connected to the transmitted coil. The coil (obtained commercially) was wound from copper wire 0.2 mm in diameter; the coil itself had an external diameter of 15.2 mm, was 10 mm in length and was wound around a ferrite core with a diameter of 13 mm. The ferrite core was topped with a 3mm flange made of the same material, with an external diameter of 16mm. The inductance at 10 kHz was 4.7 mH, as measured by a HP 4192A impedance analyzer. The currents flowing through the coil at 4.5 kHz and 9 kHz were measured to be 7.5 mA and 3.8 mA respectively.
Three methods were used to obtain and cross-check the magnetic flux density of the energised coil at its end surface: (1) by direct measurement using a Hirst GM04 Gaussmeter (Hirst Magnetic Instruments Ltd); (2) by using the Biot-Savart law directly; (3) by using the finite element modelling package, Femlab. Using these methods, the magnetic flux density at 9 kHz was estimated to be 0.031 mT, 0.021 mT and 0.03 mT respectively. These values were broadly consistent with one another and in general agreement with measured field strengths attenuated by the plate. As Figure 2 indicates, the coil was enclosed within a rectangular section mild steel box with a wall thickness of 1 mm. The box was 100 mm in height, 200 mm in width and 200 mm in length. In place of a lid, a steel plate was placed over the box, and the transmitter coil mounted at its centre. The steel plate was 2 mm in thickness and 1000 × 1300 mm in area. The dimensions of the plate were selected to be much larger than those of the transmitter coil to minimise edge effects, which dominate at higher frequencies, and to reduce coupling of the excitation and sensing coils around the edges of the plate. The receiver coil was of identical dimensions to the transmitter, but had an inductance of 11 mH. It was attached to a scanner arm on the opposite face of the plate at a height of either 3 mm or 10 mm, and aligned such that when scanned across its shorter dimension, it crossed the location of the transmitter coil. The receiver coil outputs were applied to a low-noise wide-band instrumentation amplifier (10 Hz to 4 MHz), designed with a moderate gain of 400 (52 dB). The signal produced by the amplifier was then fed to a real time DSP system.

The DSP system comprised a high-level Windows-based user interface that designed the filters, and a hardware module that executed the filters in real time. The main processing element of the DSP system was a DSP56309 (Freescale Semiconductor Inc.), operating at 100 million multiplication-accumulations per second (MMACS). This processor is a single clock-cycle per instruction cycle machine, incorporating a 24-bit, fixed-point, fractional arithmetic ALU, super Harvard architecture with 20 k-bytes of code memory space and 14 k-bytes of data memory space [6]. To take advantage of the enormous dynamic range of the ALU (144 dB), the processor was here interfaced to a Cirrus CS4271 dual-channel 24-bit sigma-delta codec sampling at 48 kHz. The DSP system was employed both to perform super-narrowband filtering and to apply high levels of digital gain with exceptional purity. The output of the DSP system was fed both to a digital oscilloscope for signal capture, and to a second, identical DSP system which performed real time spectral analysis of the signals.
2.1. Design of the filter and digital gain stage

As confirmed by the skin-depth calculations, the raw signal detected by the receiving coil was extremely weak; it was also embedded in background noise (both electronic and environmental in origin), which typically has a $1/f$ distribution. Amplification alone would not extract the signal, and so filtering was required. However, filtering would only be successful if sufficient noise bandwidth could be suppressed whilst leaving the signal bandwidth unaffected. The weaker the signal, the more specific the filter must be. Because of the extreme nature of the problem, an analogue approach would have been unacceptable, since it is virtually impossible using this methodology to design and implement filters with the required frequency precision, transition zone performance and stability. Digital filters are typically realized using either a finite impulse response (FIR) or infinite impulse response (IIR) approach. In this application, a bandwidth of 2 Hz within a frequency range of 24 kHz was demanded. Such specificity would require the kernel of an FIR filter to comprise approximately 30,000 coefficients, effectively ruling out real time operation. In contrast, IIR filters rely on recurrence formulae, where the output signal is given by

$$y[n] = \sum_{k=0}^{N} a[k] x[n-k] - \sum_{k=1}^{M} b[k] y[n-k]$$

(3)

Because of their recursive nature, IIR filters impose less computational burden on the processor than FIR types, requiring fewer coefficients to achieve the same cut-off performance. In fact, it is possible to design a low-order narrowband filter with as small a bandwidth as desired by placing the poles of the filter sufficiently close to the unit circle in the $z$-plane. In this case, a super-narrowband 2$^{nd}$ order IIR filter was implemented in software using the pole-zero placement method [6]. The difference equation is given by

$$y[n] = x[n] - 2\alpha_0 x[n-1] + \varepsilon_0 x[n-2] + 2\alpha_1 y[n-1] - \varepsilon_1 y[n-2]$$

(4)

The high level software takes as its inputs the pole and zero locations for the filter, calculates the IIR coefficients and downloads them to the DSP hardware module which then implements them in real time. Two such narrowband filters were designed, for 4.5 kHz and 9 kHz. Table 1 gives the locations of the poles; all zeros had the value zero.

After filtering, the signal was post-amplified digitally by a factor of 4096, fed to the digital -to-analogue section of the codec, and from there it was sent as a voltage to the oscilloscope. In total, the gain applied to the (filtered) signal was 1638400, or 124 dB.

3. Results

Using a motorised laboratory scanner, the receiver coil was tracked across the surface of the plate and, at 20 mm intervals, the root mean square (RMS) voltage of the processed and amplified signal was measured by the digital oscilloscope. Figure 2 depicts the results obtained from a scan in which an excitation frequency of 4.5 kHz was used, with the receiver located 10 mm above the steel surface. This distance was chosen for issues relating to the ease of scanning; however, a future objective of the work will be to detect fields at much greater stand-off distances, since the plates will in practice be embedded within concrete. The central peak in Figure 2 corresponds to the exact position of the transmitter coil. The tails at the left and right extremes of the plot are due to edge effect leakage. Figures 3 shows the data obtained using an excitation frequency of 9 kHz and a scan height of 3 mm. Although the edge effects are much stronger and the signal peak weaker, there is no question that the system is indeed detecting the signal propagated through the steel barrier.
To establish the detection limit of the instrumentation, the excitation frequency of the signal applied to the transmitter was increased and the digital filter redesigned accordingly; reconfiguring the DSP system requires only a few seconds. Specifically, measurements were taken at frequencies of 12 kHz and 13 kHz, and in both cases the signal was unambiguously recovered from the noise.

4. Discussion

The ability of the system to detect the high frequency magnetic fields is critically dependent upon both the resolution of the DSP system and the selectivity of the super-narrowband filter. At 9 kHz, the raw signal voltage produced by the receiver, when directly over the transmitter, is 0.27 \(\mu V\), RMS. Since the system can (theoretically) resolve to 1 part in \(1.678 \times 10^7\), the minute signals present in the background noise lie above its quantisation limits and can therefore be recovered and amplified. The integrated energy of the filtered signal increases as the duration of the kernel lengthens; for discrete
systems, the increase is proportional to $\sqrt{M}$, where $M$ represents the number of coefficients in the filter. The potential consequence of this relationship is significant; it implies that much higher frequencies than those discussed here could be detected, given appropriate processing algorithms and instrumentation.

The measurement instrumentation described here is not alone in its ability to detect weak magnetic fields; for example, lock-in amplifiers may be employed. The authors have compared the results obtained from this system with those provided by a Bentham 223 lock-in amplifier, which has a quoted sensitivity range of 10 $\mu$V to 1 V and a frequency range of 0.5 Hz to 100 kHz. This amplifier was used in place of the DSP system with all other instrumentation remaining in place. It was found that the amplifier was able to detect the signal up to 4.5 kHz, but not beyond. The authors appreciate that more sophisticated lock-in amplifiers are available with greater sensitivities, but cost is also a factor which mitigates in favour of the digital approach; lock in amplifiers presently cost between £1500 and £2500, whereas the digital system could be constructed for a tenth of this.

There is, moreover, a more fundamental reason why the digital filter approach is superior. The system described here can interrogate at multiple frequencies simultaneously; the filter can be designed with several super narrow pass bands, with no additional overhead with regard to processing or hardware.

In these tests, the use of a large plate enabled the edge effects to be maintained at a safe distance from the receiving coil. It is also important to emphasise that the field strengths generated in this tests were very small. The sensitivity of the receiving coil is directly proportional to its area; by increasing its radius by a factor of ten, the voltage output is increased by two orders of magnitude. The authors are now modifying the system in an attempt to detect signals extending to 20 kHz.

5. Conclusion
A system has been developed that can detect alternating magnetic fields, extending to 13 kHz, propagated through a ferrous steel boundary 2 mm in thickness.

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