Inductive pulse transmission by amplitude modulation using thin-film and electroplated microcoils

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Abstract. Inductive links are widely used for implanted biomedical applications, and with amplitude modulation their use can be expanded to the transmission of pulse trains for deep brain stimulation (DBS). Using a passive envelope detector and an integrated coil, pulse trains can be obtained with high fidelity across a load representing brain tissue. To improve the system design, a comparison is made between thin-film and electroplated coils for receiving signals. Using our inlaid electroplating process, the coil resistance can be greatly reduced, which translates to increased output levels at the load at a few megahertz. One feature of our inductive link is enhanced output from the electroplated coils at system resonance. Various rectification methods provide flexibility in obtaining desired system performance. With this technique, the implanted components for DBS could be reduced to an integrated coil and a few components. © 2005 Society of Photo-Optical Instrumentation Engineers. [DOI: 10.1117/1.1857532]

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1 Introduction

This work presents a novel method to inductively deliver pulses for electrical stimulation, for instance, deep brain stimulation (DBS), which is the application considered here. By inductively coupling and reproducing stimulation waveforms, the conventional implanted electrical stimulator can be moved outside the body. The only parts that remain inside the body would be the stimulation electrodes and a chip with integrated coils and circuitry for waveform pick-up and recovery. The advantages are less surgery, less risk of infection, flexibility with parameter adjustment, lower cost, etc.

DBS is used to treat several medical conditions, including the symptoms of Parkinson’s disease and essential tremor, such as dyskinesia, bradykinesia, tremor, and rigidity. It is implemented by delivering pulses to electrodes implanted into the thalamus, globus pallidus (GPI), or subthalamic nucleus (STN) of the brain.1–4 The pulses have widths from 20 μs to 1.5 ms, repetition rates from 2 to 2500 Hz, and amplitudes from 0.1 to 20 V.5 The precise values are determined by the response of individual patients.

In the biomedical field, DBS has conventionally been approached by implanting a signal generator inside the body.6 In the Medtronic Activa® system (Medtronic, Incorporated, Minneapolis, Minnesota) pulse generators (similar to pacemakers) are implanted into a patient’s pectoral region to provide signals.7 A wire tunnels under the skin from the chest to the top of the skull, connecting the pulse generator with the electrodes. Here, the signal generator prestores waveforms that can be selected through an inductive data link.5–10

However, researchers are still actively seeking optimum conditions for electrical stimulation. Prestoring waveforms in implanted pulse generators makes it inconvenient to adjust waveforms. This and the previously stated reasons motivate us to investigate external modulation and inductive transmission of the waveforms.

Inductive coupling is widely employed in implantable applications to power implants or transmit data. As increasing attention is directed toward downscaling of implants, microfabricated coils are gaining popularity to realize monolithic integration of implanted electronics.11–14 In those applications, integrated coils are used for power delivery, so there are no special requirements on the waveforms received at the coils. However, in applications such as DBS, wireless transmission of specific waveforms is desirable, which has not been addressed in prior reports.

Direct transmission of long-duration (suitable for DBS) waveforms through a transcutaneous transformer is difficult. Assuming $R_{\text{GEN}}$ is the equivalent internal resistance of the power supply, the inductive link (a weakly coupled transformer) has a time constant for its driving coil of $\tau = L_1 / R_{\text{GEN}}$ ($L_1$ is the inductance of the driving coil). To transmit a waveform, that time constant must be much larger than the waveform duration to avoid waveform distortion. To obtain a large time constant $\tau$, one can either try to reduce $R_{\text{GEN}}$ or use a large $L_1$. Typical values for $R_{\text{GEN}}$ are from a few ohms to several tens of ohms. If the desired time constant is comparable to or larger than that of the system with a practical power supply, it is not realistic to reduce $R_{\text{GEN}}$ alone to increase the time constant, since $R_{\text{GEN}}$ cannot be made arbitrarily small. On the other hand, because of space restrictions for implanted applications, the
receiving coils tend to have few turns, and thus small inductance \( (L_2) \). To provide usably large signals to the implants, the inductance of the driving coil is often minimized, since the output voltage scales with the effective turns ratio of the inductive link \( (\approx \sqrt{L_2/L_1}) \). As a result, the inductive link has extremely poor time-domain response when transmitting pulses of relatively long duration.

Therefore, we demonstrated an amplitude modulation scheme to transmit pulse trains for DBS through an inductive link, in which an integrated coil intercepts the modulated signal and the waveform is retrieved with high fidelity at the load. The load represents the brain tissue and is also part of an \( RC \) demodulator. Passive demodulation is adopted, which is simple, compact, reliable, and can be used for a range of carrier frequencies and power levels. Additionally, there exists a large tolerance for resistance values for a fixed capacitance at the detector, which for our application is the brain tissue of the patient. This property is important, as the brain tissue resistivity varies from patient to patient. With an amplitude modulation scheme, only integrated coils and a few components need to be implanted underneath the scalp to deliver signals for DBS.

There are prior reports on using amplitude modulation in loosely coupled inductive links for transmitting low-frequency signals to drive the gates of MOSFETs, or from an artificial heart (using amplitude shift keying) to monitoring equipment outside the body. In these cases, the coils are made from printed circuit boards (PCBs) or Litz wires, and are thus free from the limitations and parasitic effects of a microfabricated coil. Also, the volume of the signal receiving side is not a major concern in these two instances, since they are not implanted.

In comparison, our prototype minimizes the volume of the receiver side, which should be implantable upon successful development. Pick-up coils were fabricated with microtechnologies. To minimize signal attenuation arising from large internal resistance of thin-film microcoils, we have investigated techniques for integrating low-resistance coils with conventional silicon technology. Besides the inductive waveform transmission scheme, this work also presents a new inlaid electroplating method to reduce coil resistances. In this process, the coils are inlaid into the silicon substrate, and mechanical polishing can be used for planarization, thus preserving the monolithic integrability of coils with electronics. Our process shows advantages over similar procedures such as damascene and lithographic (lithography), galvanoformung (electroplating), and abformung (molding) (LIGA) processes, which is discussed in the next section. Proof of concept for inductive waveform transmission is described and a comparison of the performance of the thin-film and electroplated coils in pulse transmission is presented.

2 Coil Fabrication and Characterization

2.1 Integrated Coil Fabrication

Existing electroplating processes, such as damascene and LIGA (German acronym), are utilized to produce metal structures of large dimensions. Here we review our new electroplating process, termed “inlaid electroplating,” used to fabricate our low-resistance coils. A more complete treatment was presented elsewhere.

Both the thin-film and electroplated coils are 20-turn square spirals with a 14-mm maximum side length, and the copper lines are 80 \( \mu \)m wide on a 100-\( \mu \)m pitch. At each end of the microcoils was placed a 200\( \times \)200-\( \mu \)m bonding pad. The wafer is divided by “streets” into grids, with one coil located in each sector. The streets provide electric connections between plating contacts at the wafer edges and the outer bonding pads of the coils; hence the streets, plating contacts, etc., are etched and then filled by electroplated copper along with the coil structures. A Cu/Cr multilayer was chosen as the metallization in both cases, as the resistivity of copper is second only to that of silver. The substrates are 550-\( \mu \)-thick, 4-in. single crystal silicon (100) wafers, doped n-type with resistivity 5 to 10 \( \Omega \)-cm.

Thin-film coils were realized by e-beam evaporating the multilayer metal on oxidized wafers and then patterning by lift-off. The thin film coils consisted of a nominally 50-\( \mu \)-thick Cr adhesion layer and a 750-\( \mu \)-thick Cu layer. Electroplated coils were fabricated using the inlaid electroplating process shown in Fig. 1. First, as shown in Fig. 1(a), coil-shaped trenches were etched into silicon wafers as electroplating molds. Because of their superior mechanical strength and chemical stability, silicon molds pose few limitations on lateral and vertical geometries with mold design. In particular, silicon molding methods need not be concerned with retaining mold form and adhering well to the substrate in the aggressive electrodeposition solution throughout the process. Our molds, being composed of 14-
mm-long, 20-μm-wide, and 60-μm-high beams (or sidewalls), were found to be impossible to fabricate in conventional polymer material due to warpage of the walls. After growth of thermal silicon dioxide, a ~1-μm-thick Cu/Cr seed layer was e-beam evaporated, as shown in Fig. 1(b). For the pattern-electroplating adopted in our process, the seed layer was wet-etched away from regions other than the bottom of trenches [Fig. 1(c)], so copper grew only inside the trenches during subsequent copper electroplating, as shown in Fig. 1(d). When spinning photoresist to protect the metallization at the trench bottoms, 3-D topologies and centrifugal force cause the photoresist to deplete at some sites. The rough surface of exposed silicon at the bottom of the trenches further worsens the photoresist coverage. The problem was resolved by using high-viscosity photoresist AZ 4620P spun at 2000 rpm. With this spin rate, AZ 4620P can be coated sufficiently thick at all sites without obvious edge beads. Corresponding to Fig. 1(d), copper was then electroplated inside the trenches. Because patterned electroplating is adopted in our process and the electrical path to inside turns must trace through all the coil lines, there is an electrical potential drop between the two coil ends arising from the resistance of the seed layer (a few hundred ohms). Consequently, the problem of plating uniformity is more severe with our coil design. Measures were taken to improve plating uniformity, including depositing a thicker seed layer, lowering the plating current, using the "streets" as current "thieving" rings, and adopting reverse pulse plating, which serves to reduce nonuniformities in plated coils. Figure 2(a) shows the uniformity of electroplated coil inspected in a scanning electron microscope (SEM). The electroplated copper is reasonably flat and well contained within the sidewalls. To obtain a leveled surface, copper is overplated above the wafer [Fig. 1(e)] and polished back [Fig. 1(f)]. Due to equipment limitations, polishing was performed with lapping films on a rotating wheel without chemical slurries. An SEM micrograph of a polished coil is shown in Fig. 2(b). By surface profiling, an approximately 70-nm step is observed between the outermost turn and Si surface. The surface roughness is acceptable for subsequent photolithography steps. The thickness of the electroplated coils is around 44 μm, which is determined by the depth of the trenches as measured at the bonding pads across the wafer.

The inlaid electroplating process bears a resemblance to the so-called HEXSIL process developed at the University of California Berkeley, in that both processes adopt deep trenches etched in silicon wafers as molds. However, HEXSIL is a template replication process for stand-alone polysilicon structures, whereas inlaid electroplating is used to obtain metal microstructures. Inlaid electroplating also yields advantages compared with the aforementioned damascene and LIGA processes. Copper damascene is already adopted for interconnect fabrication by many IC manufacturers. However, in the damascene process the copper seed layer is not patterned prior to electroplating, and polishing is required afterward. The plating thickness is thus limited to several microns, so as not to impose too great a task to polishing. Metal microstructures of large thicknesses are typically obtained through LIGA processing, using either x-ray or UV light. However, LIGA falls short when integrating metal structures with other microde-
count for coil self-inductance and dc resistance, and a capacitor in parallel accounts for capacitive coupling between the copper lines and the conductive substrate.

The inductance of a spiral integrated coil can be calculated as:

\[
L = 10^{-7} \frac{(OD + ID) \cdot (OD - ID)^2}{p^2} \left\{ \ln \frac{OD + ID}{OD - ID} + 0.2235 \frac{OD - ID}{OD + ID} + 0.726 \right\}
\]

where \( L \) is in Henries, \( p \) is the coil pitch in meters, and \( OD \) and \( ID \) are the outer and inner diameters of the coil in meters, respectively. From Eq. (1), it can be seen that coil inductance increases with its area and turn number, which are functions of \( p \), \( ID \), and \( OD \). Also, the coil inductance has no dependency on its thickness. With \( p = 100 \, \mu m \), \( OD = 14 \, mm \), and \( ID = 10 \, mm \) for 20 turns, \( L \) is calculated as 9.81 \( \mu H \) for 20-turn coils. Coil resistances are calculated using \( R = \rho l/S \), where \( \rho \) is copper resistivity, and \( l \) and \( S \) are the length and cross sectional area of the metal line, respectively. Bulk copper resistivity is 1.67 \( \mu \Omega \)-cm, and
that of electroplated copper is slightly higher, between 1.94 and 2.05 μΩ-cm. Using an average side length of 12 mm and a cross sectional area of 0.8 μm² for thin-film coils and 44 μm² for electroplated coils (including the 50-μm-thick adhesion layer for both), their resistances were calculated to be 250.5 Ω for thin-film coils and 5.45 Ω for electroplated coils. Because the substrate is floating, conventional analysis of capacitance does not apply, so the coil capacitance is not determined analytically. The calculated values are compared with measurements later.

Figures 3(b) and 3(c) show the measured impedance of thin-film and electroplated coils, respectively. Using curve fitting, the extracted model elements are: \( R = 275.8 \) Ω, \( L = 9.57 \) μH, and \( C = 114.8 \) pF for thin-film coils, and \( R = 5.95 \) Ω, \( L = 9.57 \) μH, and \( C = 243.6 \) pF for electroplated coils. As expected, with increased thickness, resistance goes down and parasitic capacitance goes up. The coil inductance has no significant dependence on the metal thickness.

The measured coil parameters have good agreement with the calculated values, which supports the usefulness of Eq. (1) in designing microcoil inductance. The slight differences between the measured and calculated resistance may be attributed to the adhesion layer, estimate of coil thickness, or variations in the densities of metals. Nevertheless, the close agreement suggests that the electroplated copper is void-free.

3 Inductive Link with an Integrated Coil

For transcutaneous electrical stimulations, signals are transmitted by an inductive link from an external driving coil to a receiving coil inside the body. The coupling coefficient of the inductive link depends on the distance and alignment between the driving and receiving coils. In our DBS scheme, the receiving coil would be placed beneath the scalp, so the separation between two coils should be approximately the thickness of the scalp and hair. Joung and Cho\(^{28}\) reported that a 1-cm spacing between their two coils yielded \( k = 0.4 \) for a transcutaneous transformer. The test transformer used to characterize our integrated receiving coils consists of a driving coil wound around a ferrite core with a 2-mm airgap, which can provide a coupling coefficient \( k \) up to 0.8, depending on the separation between the driving and receiving coils. To simulate an implanted application, we chose a coupling coefficient \( k \) of approximately 0.4.

Conventionally, \( k \) is determined by measuring the inductance of (looking into) one coil with the other coil first open- and then short-circuit. However, when integrated coils are used in an inductive link, this method may give erroneous results because of coil parasitics. Figure 4 is a schematic of an inductive link with an integrated coil. The following relations can be obtained at the coil terminals:

\[
V_{L1} = j\omega L_1 I_1 + j\omega M I_2 \tag{2}
\]
\[
V_{L2} = j\omega M I_1 + j\omega L_2 I_2 \tag{3}
\]

where \( M = k\sqrt{L_1 L_2} \) is the mutual inductance. When coil 1 is open, \( I_1 = 0 \) and \( V_{L2} = j\omega L_2 I_2 \). The measured inductance from the coil-2 side should be \( L_2 \). When coil 1 is short-circuited, \( V_{L1} = 0 \), and \( V_{L2} = j\omega (1 - k^2) L_2 I_2 \). Then the measured inductance, now the leakage inductance, is \( L_{\text{leak}} = (1 - k^2) L_2 \). Therefore, the coupling coefficient can be determined from the difference between the measurements of coil inductances. However, because of parasitic capacitances, the measured coil impedances include capacitive effects, and using those values to calculate \( k \) directly can be misleading. It was shown in our previous work\(^{28}\) that direct calculation yields frequency-dependent \( k \), that is, it initially increases slightly with frequency then drops suddenly beyond a certain frequency. It is not that \( k \) changes with frequency, but that direct calculation does not de-embed the parasitics of coil 2. So here, similar to the coil characterization in Sec. 2.2, coil inductances are extracted by curve fitting using the coil model of Fig. 3(a). Measurements were performed looking into coil 2, as shown in Fig. 4, since in this configuration it is easier to extract the parasitics. \( L_2 \), the coil inductance in the air gap with coil 1 open,
is found to be 24.03 \mu H; then, short-circuiting coil 1, the extracted leakage inductance of coil 2 is 19.89 \mu H. Therefore, \( k \) is calculated to be 0.415.

A less-than-unity coupling coefficient, which is equivalent to finite leakage inductance, gives rise to a frequency-dependent transfer function for an inductive link. Combined with the capacitance on the receiving side, the inductive link reaches its maximum voltage gain when the capacitance and leakage inductance combination comes into resonance. This property has been exploited to obtain additional voltage gain in this work and in Refs. 31 and 32.

The open circuit voltage gain of an ideal inductive link is \( k \sqrt{L_2/L_1} \). Typically, \( L_2 \) is limited in implanted applications due to space constraints. To achieve a usable output level, \( L_1 \) is also limited. In our experiments, \( L_1 = 162 \mu H \); as a result, the coil driven by a standard 50-\Omega source has a time constant of 3.25 \mu s. Because pulse trains for DBS are typically 50 to 90 \mu s, much longer than the coils’ time constant, waveforms can be distorted when transmitted directly through the transformer. It is not practical to drive these circuits with small source impedances to

![Fig. 7 Oscillographs of input, carrier, and output signals from electroplated microcoils for 200-, 700-, and 1200-\Omega loads at a carrier frequency of 2.6 MHz.](image-url)
preserve the waveforms in this manner. Therefore, amplitude modulation is used to transmit long-duration pulse trains for our application.

4 Pulse Transmission by Amplitude Modulation

The amplitude modulation scheme for transmitting pulse trains is shown in Fig. 5. Signal demodulation is realized by a passive envelope detector, composed of a rectifier and low-pass RC filter. This type of envelope detector can be used with a wide range of carrier frequencies. Also, there exists a large tolerance for resistance values with a fixed capacitance in the detector. These properties are useful for our proposed application of DBS, because the resistance in the detector represents that of the brain tissue, which varies among patients.

The average resistivities of brain tissue are about 350 Ω·cm for gray matter and 390 Ω·cm for white matter, but range from 200 to 1200 Ω·cm. Based on these average resistivities and data from Medtronic, Incorporated specifying a load resistance of 1 kΩ for the Activa® DBS system, we used resistances of 200, 700, and 1200 Ω in our experiments. The lower resistances represent worst cases in terms of output voltage. Operation at higher resistances poses no additional difficulties, including rise and fall times of the waveform.

In the experiments, a pulse width of 90 μs was used as a representative value. The carrier signal generated by an HP 33120A was modulated by a Wavetek model 19 function generator. A bridge rectifier of four 1-N 4148 diodes and a capacitance of 4.5 nF were used for demodulation. The capacitor is large for conventional CMOS circuits, but it is acceptable for our application. Using a parallel-plate configuration with 50-nm SiO₂ as a dielectric layer, the capacitor takes an area of 6.5 mm², which poses no problems for integration with the coil.

Output levels from the three resistances were measured over the frequency range from 300 kHz to 5 MHz, as shown in Fig. 6, for both thin-film and electroplated coils at 20 Vpp input. With electroplated coils, higher outputs were obtained for all three loads. At resonance, the outputs increase by a factor of about 4, and the peaks are more pronounced than with thin-film coils. This is due primarily to the smaller resistance of the electroplated coils. Also, the system resonates at a lower frequency, which is due to the larger parasitic capacitances of the electroplated coils.

The oscillographs of output with the electroplated coil are shown in Fig. 7 for loads of 200, 700, and 1200 Ω. In the figures, Vp-p(1) represents the input modulated signals and Vp-p(2) represents the output levels. Though the carrier frequency and load resistance change dramatically, well-shaped pulse trains are obtained at the output. The rise and fall times are longer for larger resistance loads, but they can be adjusted through the filtering capacitor.

As can be seen in Fig. 6, smaller loads lead to lower resonant frequencies. In power electronics, it is common practice to represent the rectifier and filter with an equivalent ac resistance. The equivalent ac resistance is calculated by the power dissipation in the receiving coil. For bridge rectification, the equivalent AC resistance is expressed as

$$R_{equ} = \frac{R_{load}}{2} \left( 1 + 2 \cdot \frac{V_{diode}}{V_{load}} \right)^{1/2}$$

where $R_{load}$ is the load resistance, and $V_{diode}$ and $V_{load}$ are the rms voltages across the diodes and the load, respectively. The second term is the equivalent resistance of the diodes. Because diodes have nonlinear I-V characteristics, the equivalent ac resistance is smaller at larger diode currents. This is called "rectifier regulation." Figure 8 shows the output versus frequency at different input levels for electroplated coils with 200-Ω load with bridge rectification. It can be seen that the resonant frequency shifts to lower values when the input levels increase and the ac resistance of the diodes decreases, as the resonant frequency can be roughly expressed as

$$f = \frac{1}{2 \pi} \left( \frac{1}{L_{tot}C_{sec}} - \frac{1}{2R_{equ}^2C_{sec}^2} \right)^{1/2}$$

The full-wave rectifier can be replaced by a single diode to reduce the component count. The penalty is increased ripple of the output at low frequency (below 1 MHz). However, the rectification consumes only one diode voltage drop instead of two, so the output level is slightly higher.

To achieve higher output at frequencies below resonance, a voltage doubler can be used in place of the full-wave rectifier, as for example, the circuit shown in Fig. 9. The output levels from the three loads were measured again from 300 kHz to 3 MHz for rectification with voltage doubling. The results for both thin-film and electroplated coils are shown in Fig. 10. Below resonance, the output voltages
outputs of thin-film and electroplated coils with voltage doubling.

The receiving coil was fabricated by microelectronic technology, so it is expected that the detector circuitry can be transmitted with high fidelity through such an inductive link. Thus, only the receiving coil in lieu of fabricating such integrated components as large capacitors.

5 Conclusion

Inductive links are used extensively in implanted biomedical applications, but their usage is limited when transmitting low-frequency pulse trains because of waveform distortion during direct transmission. However, with amplitude modulation, it is demonstrated that pulse trains intended for DBS can be transmitted with high fidelity through such an inductive link. Thus, only the receiving coil need be implanted into a patient’s body.

An integrated coil as the receiving coil can minimize the size of the implants. In this work, the properties of a loosely coupled inductive link are studied with thin-film and electroplated microcoils, respectively, as the receiving coil. The parasitic capacitance of an integrated coil together with leakage inductance of the link gives rise to an output peak at the system resonance frequency. For electroplated coils, higher output is obtained over the measured frequency range.

Two types of rectification methods are used for the demodulation of signals, a bridge rectifier and a voltage doubler. They exhibit different frequency-dependent properties, which can be taken advantage of according to the needs of the signal transmission system. Because fabrication of the receiving coil is compatible with standard microelectronics technology, integration of coils with voltage regulation or other control circuitry may be possible in the future.

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