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Optimization Possibilities for DC Current Transformer

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Abstract— The paper analyzes the consumption performance of DC current transformers based on linear class B, and half bridge class D compensation amplifiers, as well as self-oscillating flux gate current transformers based on push-pull and source-sink output stages. Compensation amplifiers, push-pull and source-sink output stages are used to generate feedback current in order to compensate magnetic flux in the magnetic concentrator core, produced by electrical current flowing through a conductor under measurement. The focus of the analysis is the investigation of conditions for appearance of the bus-pumping effect to which all switching versions of DC current transformers are prone. All sufficient conditions for bus pumping effect appearance are explained for the most critical case of DC current measurement; the main issues are analyzed, giving better insight to power consumption, and possible hazards to DC current transformer circuitry. It is highlighted the existence of particular exploitation conditions that lead to the state where the class D based DC transformers are inferior to conventional class B based counterparts. Finally, when a linear compensating amplifier is replaced with a switching counterpart, it has been demonstrated that it is possible to expect a feasible improvement in energy consumption in the middle of the ideal half-class D class and classical B class realisations.

Keywords- current measurement, DC current transformer, Hall sensor.

I. INTRODUCTION

Modern technical systems show an ever-growing trend of demand for current measurement, which must have high precision, wide bandwidth and comply with a broad range of safety standards. There are many kinds of suitable current transducers, offered on the electronics component market, that are used in a variety of applications, ranging from classical industry measurements up to ultra-precise measurements in the nuclear accelerators[1]Error! Reference source not found.[2][3][4][5]. On the other hand, there is a growing tendency for sensor fusion in the industrial Internet of Things (IoT), what requires energy efficient sensing, especially for the case of distributed current sensing applications [7].

De-facto industrial standard for AC + DC current measurement, with galvanic isolation whose clearance can satisfy the most demanding safety standards, are current transducers based on a flux concentrator, and current measurement are preformed indirectly by measurement of the concentrated magnetic field. The flux concentrator is constructed as a high permeability magnetic core that focuses magnetic flux lines, which are generated by electrical current flowing through a conductor, (Fig.1). Therefore, insulation and distance from the conductor to the core can be optimized to withstand arbitrarily large breakdown voltages, which make

such transducers ideal for many industrial and energy applications.

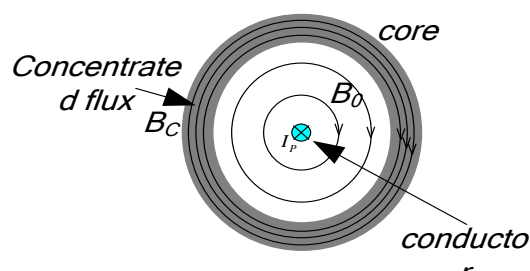


Fig. 1 Magnetic flux density inside soft magnetic material of the concentrator core is $B_c = \mu_r B_0$, where μ_r is relative permeability of core magnetic material, typically greater than 1000

The main obstacle in IoT application area of such transducers is accuracy – power efficiency trade-off. In order to address mentioned obstacle, several low power versions of flux concentrator based transducers, based on class D half –bridge switching topology, are proposed recently [8][9][10].

Unfortunately there are several contradictory design issues which should be seriously addressed, in order to correctly adjust this type of transducer to low power regime of operation. In this paper, practical limitations of using class D based topology were analyzed, as well as an illustration of the negative effects of the straightforward replacement of linear amplifiers with their class D equivalents.

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Nomenclature used in the paper assumes that DC quantities are represented by uppercase symbols with uppercase subscripts, for example V_{OUT} . AC quantities are represented by lowercase symbols with lowercase subscripts, for example v_{out} . Total quantities, AC+DC are represented by a lowercase symbols with uppercase subscripts, for example $v_{OUT} = v_{out} + V_{OUT}$.

II. CURRENT TRANSDUCERS BASED ON MAGNETIC FLUX CONCENTRATOR

Concentrated magnetic field can be measured with some sort of magnetic sensor, typically Hall element [3][2][11]. Also, many other kinds of magnetic sensors can be successfully used [2]. In that case the system consists of a ferromagnetic concentrator that surrounds the current conductor and is designed with a little window, called the air gap, into which the magnetic sensor IC is placed, (Fig.2). Output signal of the transducer is magnetic sensor's output signal, amplified by an operational amplifier, proportional to the magnetic flux density inside the core gap. Since the magnetic flux density linearly depends on the current I_p , output signal is measure of the same current $V_o = k \cdot I_p$, where k is a constant of proportionality.

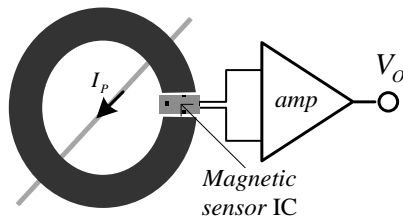


Fig. 2 Open-loop Hall effect current transducer based on flux concentrator and magnetic sensor. Output signal is sensor output voltage amplified by operational amplifier

Such a kind of current transducers are offered to the market from many vendors, for example [3], where they are named as Open loop Hall effect current transducers. The advantages of the transducer include low cost, small size, lightweight and especially low power consumption, what is important when measuring high currents, typically greater than 300 A [3]. As with most magnetic based measurement techniques, insertion losses are very low. Primary current overloads can be easily handled although it may result in some magnetization of the core creating an offset shift, called remanence, or magnetic offset. Compared to other technologies the limitations of open loop transducers are smaller accuracy, moderate bandwidth and response time, a larger gain drift with temperature, and a limitation on the current frequency product (power bandwidth). In many applications the advantages outweigh the limitations and an open loop solution is advised, especially in battery-powered circuits due to their low operating power requirements.

Compared to the open loop transducer just discussed, Hall effect closed loop transducers (also called Hall effect 'compensated' or 'zero flux' transducers) have a compensation circuit that significantly improves performance [3]. Alternatively, principle of magnetic flux sensing can be based on some other physical phenomena.

In the "zero-flux" method, which is based on a negative feedback circuit that includes a magnetic circuit as shown in the (Fig.3), the primary conductor whose current I_p is to be measured is inserted through the hole of a toroidal core, or any similar type of core (Fig. 3). Compensation current I_s is passed to through a secondary coil so it cancels the magnetic flux produced in the core by the current under measurement. This method has the advantage of compensating the effects of the nonlinearity of the given magnetic material, produced by the operational magnetic flux, and keeps it at a very low level. A typical sensor design of this type requires the addition of an external resistor (shunt resistor R_s) defined by the user, which is connected to the ground at one end, providing the similar operation as the standard AC current transformer. For that reason generic name for all such transducers is a DC Current Transformer (DCT) [1].

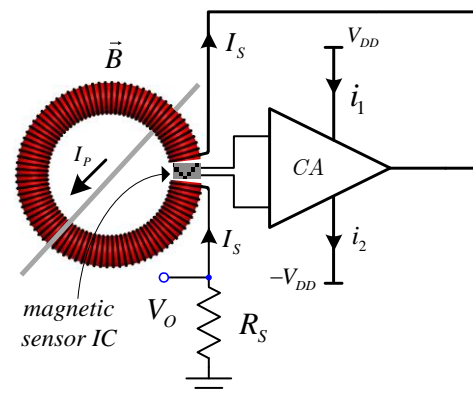


Fig. 3 DC current transformers (DCT) - basic working principle. Magnetic sensor can be Hall generator, flux-gate sensor, GMR sensor, etc

Concentrated magnetic field can be detected using different types of sensing element, placed in the gap of the ferromagnetic core. Apart to Hall generator as a magnetic field sensing element, fluxgate sensor, or GMR [12] sensor can be used, as well as some more magnetic sensor types.

The presence of the air gap in the magnetic concentrator core makes the transducer sensitive to external magnetic fields, and without proper shielding, measurement accuracy can be dramatically reduced. Additionally, using a Hall element as the magnetic field detector, an offset voltage and the associated drift usually occurs.

Perfect alternative to mentioned DCTs, is a second harmonic fluxgate DCT with ungapped magnetic core, but it is often too expensive and complex to fit economical requirements of final application [3].

A reasonable compromise is a self-oscillating fluxgate current sensor [9][10][17]. It is low cost time-domain alternative to second harmonic type with several advantages. Basically, (Fig. 4), it is an oscillator with current transformer with a saturable core as a nonlinear reactor [17]. The primary winding of used transformer T is a single conductor through the core opening, whereas a secondary winding with n turns, together with a burden resistor R_s , forms nonlinear RL circuit. It is shown [17] that average value of the v_s is equal to $\alpha \cdot I_p$, where α is a stable parameter whose stability weakly depends on the core magnetization curve, secondary winding resistivity and comparator output impedance.

Even in this basic configuration, accuracy and linearity exhibits almost the same accuracy and linearity compared to closed-loop DCTs based on Hall Element [17][18]. Where greater accuracy or better handling of output signal are required, double feedback structures are proposed [8][9][10].

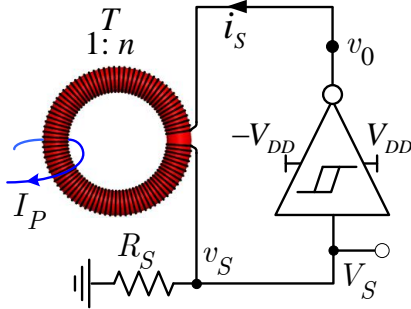


Fig. 4. Basic self-oscillating fluxgate current sensor

III. POWER CONSUMPTION IN DC CURRENT TRANSFORMERS

Common part for all DCTs is compensation amplifier (CA), together with secondary winding N_S and shunt resistor R_S . The voltage equivalent of magnetic induction B is amplified by a compensation amplifier CA, which generates current through a secondary winding I_S . This current is used for compensation of magnetic flux in the core (Fig. 3). If the number of turns in secondary winding is N_S , the secondary current will be $I_S = I_P / N_S$ what is equivalent as in the case of AC current transformer.

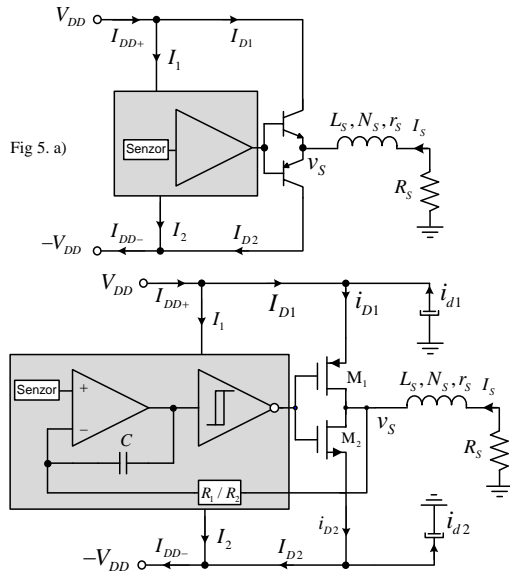


Fig. 5 Compensation amplifiers: a) linear amplifier with output level in class B and b) switch amplifier with half-bridge output in class D.

Unlike AC current transformer, which is a passive circuit in the basic realization, and which measures only AC component of the I_P , DCT is always an active circuit and measures DC + AC component of the current I_P . CA usually needs bipolar power supply, for example: $\pm V_{DD} = \pm 12$ V.

Although, the current I_S is smaller N_S times than the primary current I_P , it cannot be called a “small”, especially

when we refer to sensors based on modern low-power electronics. For the sake of illustration, if the measured current has a maximum value of 100 A, and if $N_S = 1000$ then the maximum value of the secondary current is 100 mA. If CA is linear amplifier (class B or AB), then I_S goes directly through the power supply and consequently affects the power consumption with $P_D = V_{DD} \cdot I_S = 1.2$ W what is too much for battery powered IoT transducer.

Since DCT with linear CA is the dominant configuration on the market, alternative, low power intended solutions are proposed in technical literature [8]-[10], where used CA is a switching, class D half-bridge amplifier. However, the aforementioned literature does not offer a full treatment of possible improvements and problems in efficiency, compared to the linear CA solutions.

IV. POWER CONSUMPTION IN DC CURRENT TRANSFORMERS

In the following analysis it is considered that I_P and I_S are positive currents with reference directions as denoted in (Fig. 3), so that all discussions refer to the DC consumption.

The principal schematic, for all concentrator gapped DCTs, is shown in (Fig. 3). Regardless of the type of flux detector, the common component for all DCTs is the CA, along with the secondary winding N_S and shunt resistor R_S . DCT power consumption P_D consists of a standby part P_0 , related to standby current I_{C0} , and a variable part P_V related to I_S .

$$P_D(I_{C0}, I_S) = P_0(I_{C0}) + P_V(I_S)$$

Standby part of the power consumption can be easily reduced to acceptable levels compatible to low power requirements, applying a proper design of transducer electronic. On the other hand, variable part of the consumption depends on secondary current I_S , which has a value independent of the realization of the transducer's electronics. So the variable part of consumption it critical target for transducer low power improvement.

In order to perform fair-enough comparison in further discussions, let us consider two DCTs that differ only in CA. The first transducer (IT) (Fig. 5a) is based on a linear CA, whereas the second (SWT) is based on a class-D half-bridge CA (Fig. 5b). All other circuitry is the same for both transducers, with topology presented in Fig 3. We assume that power supplies are $\pm V_{DD}$ and are generated by unidirectional sources. Primary current is I_P , and compensation current is I_S . For both DCTs compensation winding inductance and resistance are L_S and r_S respectively. In addition, we neglected resistance of the closed switches M_1 and M_2 , as well as switching losses.

In a case of IT the total power consumption $P_D = P_{DL}$ consists of a standby part $P_0 = P_{0L}$ and a variable part $P_V = V_{DD} \cdot I_S$:

$$P_{DL}(I_S) = P_{0L} + I_S \cdot V_{DD} \quad (2)$$

Such dependence could be verified in literature [11] (LEM IT 600-S). Equation (2) can be considered the same for all allowable R_S values.

Block diagram of SWT is displayed in Fig 5b. For example,

in a case of Premo DCT 700A, power consumption dependence is in the form $I_{CC} = 200 \text{ mA} + f(I_S)$ [13]. Constant part represents internal consumption described by two standby currents I_1 and I_2 , giving corresponding standby power consumption:

$$P_0 = P_{0S} = V_{DD} (I_1 + I_2). \quad (3)$$

It is worth to mention that for the case of IT, standby current of linear CA can be made extremely low by using low power operational amplifier, enhanced with BJT or MOS class B output stage, which in the case of switching CA is not so easy. Since all other circuitry in both DCTs are the same, it is easy to achieve that $P_{0L} \leq P_{0S}$. However, for the first discussion, we can assume $P_{0L} = P_{0S}$, as well as $I_1 = I_2 = I_0$.

We can express power consumption of SWT as:

$$P_{DS}(I_S) = P_{0S} + (R_S + r_S) I_S^2 = P_{0S} + I_S \cdot ((R_S + r_S) I_S). \quad (4)$$

Since condition $|V_S| = |(R_S + r_S) \cdot I_S| < V_{DD}$, must be satisfied for proper operation of the circuits from Fig 5, and if $P_{0L} = P_{0S}$, then it looks like that $P_{DL} > P_{DS}$ stands for any operation condition, and main conclusion from the original paper looks correct [8]. Graphical representation of described consumptions is calculated using (2), (3), (4), and depicted on Fig. 6.

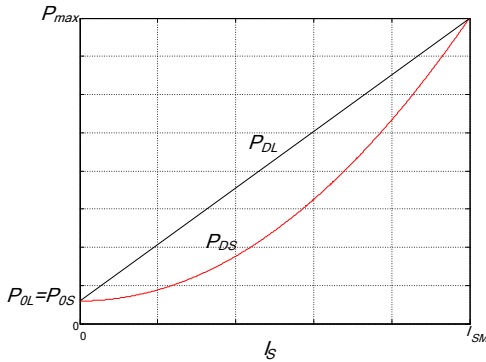


Fig. 6 Power consumption of two transducer types IT and SWT. Standby consumptions are the same, $P_{0L} = P_{0S}$, as well as other design parameters

In order to justify low power possibilities of SWT authors in [8] compared their design DCT-700A with a single device from the market, LEM IT 600-S, and only for a single condition (paragraph 4. section D in [8]). They left a question concerning power efficiency: is it really improved for the whole range of the external power supply (10 V-30 V), and the whole range of the burden resistor value ($1 \Omega \leq R_S \leq 20 \Omega$)?

To answer that question, let us reexamine the design proposed in [8]-[10].

V. POWER FLOW IN CLOSED LOOP DCTs

Since L_S is sufficiently large, we can assume that compensation current i_S is a pure DC, $i_S = I_S$. During first part of a switching period T , switch M_1 is turned on DT seconds, where D is duty ratio of CA, $0 \leq D \leq 1$. During the second part of the switching period, switch M_2 is turned on $(1-D)T$

seconds. Depending on D , current I_S is equal:

$$I_S = \frac{-V_S}{R_S + r_S} = \frac{1}{R_S + r_S} \left(\frac{(1-D)T \cdot V_{DD} - DT \cdot V_{DD}}{T} \right) = (1-2D) \cdot \frac{V_{DD}}{R_S + r_S} = (1-2D) I_{SM} = \left(\frac{1}{2} - D \right) \cdot 2 I_{SM} \quad (5)$$

where parameter

$$I_{SM} = V_{DD} / (R_S + r_S) \quad (6)$$

is theoretically maximal possible I_S for used R_S :

$$I_S(D) = \begin{cases} I_{SM}, & D = 0 \\ -I_{SM}, & D = 1 \end{cases} \quad (7)$$

Thus, compensating feedback loop adjusts D to the value:

$$D = 0.5 - I_S / (2 I_{SM}). \quad (8)$$

It is obvious that for $I_S = 0 \Rightarrow D = 0.5$.

DC component of i_{D1} , Fig 5b, is equal to $I_{D1} = -DI_S$ while the DC component of i_{D2} is equal to $I_{D2} = (1-D)I_S$. According to Fig. 5b and equation (6):

$$\begin{aligned} I_{D1} &= -DI_S = -0.5 I_S + I_S^2 / (2 I_{SM}), \\ I_{D2} &= (1-D) I_S = 0.5 I_S + I_S^2 / (2 I_{SM}). \end{aligned} \quad (9)$$

The minimum of $I_{D1}(I_S)$ is equal to $-I_{SM}/8$ and reached for $I_S = I_{SM}/2$ and $D = 1/4$. Normalized $I_{D1}(I_S)$ and $I_{D2}(I_S)$ for $0 < I_S < I_{SM}$ are presented in Fig. 7. It is obvious that for $0 < I_S < I_{SM}$, $I_{D1}(I_S)$ is negative, while $I_{D2}(I_S)$ is positive, and that positive supply bus, at the worst case, must absorb extreme value of I_{D1} :

$$I_{D1}(I_{SM}/2) = -I_{SM}/8$$

For $-I_{SM} < I_S < 0$ graphic on Fig. 7 is axially symmetric.

For the case of ideal power supply, energy conservation lays in fact that negative $I_{D1} = -DI_S$ returns part of secondary winding energy to positive supply:

$$P_{tot} = P_Q + \underbrace{V_{DD} I_{D1}}_{\text{negative}} + V_{SS} I_{D2}$$

But in real case, if we carefully look at the Fig. 7, problem can be easily noticed: $I_{D1}(I_S)$ is negative and since power supply in proposed transducer [8] is unidirectional, there is no possibility for bidirectional power flow. Therefore, supply currents I_{DD+} and I_{DD-} must not be negative, meaning that following inequalities must be satisfied:

$$I_{D1} + I_1 \geq 0, I_{D2} + I_2 \geq 0. \quad (10)$$

If the internal SWT dissipation, related to I_1 and I_2 , is not large enough, problem occurs, since unidirectional power supply is incapable to absorb negative I_{D1} or I_{D2} . The problem is similar to well-known effect in theory of audio amplifiers, named as buss pumping effect [15]. In that

situation, power bus with negative supply current increases voltage level, increasing in a same time internal consumptions inside of the SWT, in order to dissipate energy returned from the secondary winding. Repercussions of this effect can be diverse, but all of them are negative. In one of the possible scenarios, it is possible to damage by overvoltage both SWT and other components that are on the same power bus. Contrary to SWT, such effect does not exist in IT.

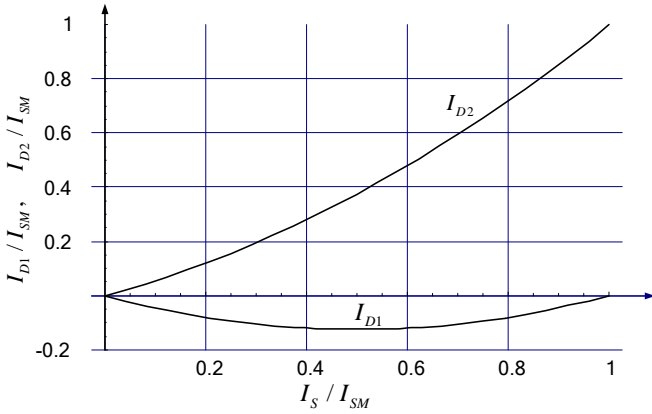


Fig. 7. Normalized $I_{D1}(I_S)$ and $I_{D2}(I_S)$ for $0 < I_S < I_{SM}$.

VI. STANDBY POWER CONSUMPTION AND PROPOSED SWT WORKING CAPABILITIES

According to previous section, SWT with small standby power consumption can face some substantial restrictions in working capabilities. These restrictions can be defined against values of I_1 and I_2 . For $I_S > 0$ it should be satisfied that:

$$I_{D1} + I_1 = -DI_S + I_1 \geq 0 \Rightarrow DI_S \leq I_1. \quad (11)$$

Taking into account (9), I_{D1} reaches minimum for $I_S = I_{SM} / 2$, as can be seen in Fig 7. When we use (8) and (11), substituting I_S with $I_{SM} / 2$, we get inequality: $I_{SM} / 8 \leq I_1$. For $I_S < 0$, the same inequality can be derived for I_2 : $I_{SM} / 8 \leq I_2$.

If measurement range $I_S \in (-I_{Smin}, +I_{Smax})$ includes $\pm I_{SM} / 2$, i.e. $-I_{Smin} < -I_{SM} / 2$ and $I_{SM} / 2 < I_{Smax}$, assuming $I_2 = I_1 = I_0$, we can get necessary condition that enables SWT to absorb negative I_{D1} or I_{D2} :

$$I_{SM} / 8 \leq I_0 \quad (12)$$

Now, we can take closer look on SWT proposed in [8].

According to data sheet 0, for external single supply of $V_{CC} = 12V$, current consumption is $I_{CC} = 200 \text{ mA} + f(I_S)$. Using $I_1 = I_2 = I_0$, internal standby current can be calculated:

$$P_{0S} = 200 \text{ mA} \cdot 12 \text{ V} = 2I_0 \cdot 12 \text{ V} \Rightarrow I_0 = 100 \text{ mA} \quad (13)$$

On the other hand, compensation winding maximum resistance is $r_s = 10 \Omega$, and burden resistance can be $R_S = 1 \Omega$, giving possible $I_{SM} / 8$:

$$I_{SM} / 8 = (12 \text{ V} / 11 \Omega) / 8 = 136 \text{ mA} > I_0! \quad (14)$$

The most critical situation is for $I_S = I_{SM} / 2 = 544 \text{ mA}$, what belongs to measurement range.

Similar conclusion can be derived if standby power consumption is taken from graphs [13], or from [8].

VII. PRACTICAL REPERCUSSIONS

Low power design is well-established knowledge, we offer reference [14], Texas Instruments control IC for Vacuumschmelze sensors with negligible standby current. Complete control electronic from [8] can be designed in low power fashion, including low current LED (1 mA), and relay of latching type.

In order to present extreme, but illustrative theoretical example, let us apply concepts from [8], [9] on the design procedure for two previously assumed DCTs that differ only in CA. We want to achieve true low power consumption with target standby current of $I_0 = 15 \text{ mA}$. It is small value, but not far distant from reality [14]. Other specifications are: $\pm V_{DD} = \pm 12 \text{ V}$, $N_S = 1000$, $r_s = 10 \Omega$, $0 \leq R_S \leq 10 \Omega$, and $I_{Pmax} = 700 \text{ A}$.

In a case of SWT, maximal I_{SM} is achieved for $R_S = 0$ and is equal to $V_{DD} / r_s = 1.2 \text{ A}$. $I_{SM} / 2 = 600 \text{ mA}$ what belongs to measurement range. Since (12) must be satisfied, we have to artificially increase I_0 , or to decrease I_{SM} by adding serial resistance to R_S . However, by adding serial resistance, measurement range will be decreased, and that is not an acceptable option. Therefore, the only option is to artificially increase I_0 to the value of $150 \text{ mA} = I_{SM} / 8$, and $P_{0S} = 2V_{DD} \cdot 150 \text{ mA} = 3.6 \text{ W}$.

On the other hand, if we design an IT with the same characteristics, there is no need for increasing of I_0 . Standby power dissipation can be calculated using (1) as $P_{0L} = 2V_{DD} \cdot 15 \text{ mA} = 0.36 \text{ W}$.

In order to discuss potential SWT advantage over IT, power consumption ratio ρ is calculated for $0 \leq R_S \leq 10 \Omega$, and displayed in Fig. 8.

$$\rho = (P_{DL} - P_{DS}) / P_{DL} = 1 - \frac{P_{0S} + (r_s + R_S) I_S^2}{P_{0L} + V_{DD} \cdot I_S} \quad (15)$$

It can be seen that SWT has smaller power consumption only for $R_S = 0$ and $I_S > 0.4 \text{ A}$, we can conclude that for defined set of electrical specifications, and achieved I_0 , the IT is much better than the SWT.

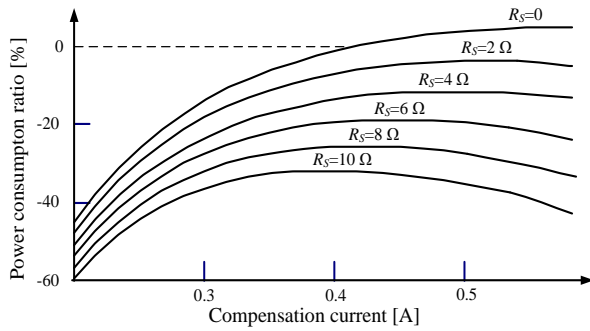


Fig. 8. Power consumption ratio of the compared transducers versus the primary current value, for $0 < R_S < 10 \Omega$.

It is important to note that the completely different requirements can be adopted, where the situation would be completely opposite, but such an analysis would not be useful because it does not reveal the essence of the problem.

VIII. CASE OF THE SINGLE FEEDBACK SELF OSCILLATING FLUX-GATE DCT

Self oscillating single feedback DCT realization, proposed for example in [17][17], can be designed using a Schmitt trigger with class B output stage. Schematic of the complete DCT is depicted on Fig. 9. It is obvious that power consumption in that case has same characteristics as in case of classical IT: complete secondary current flows directly through the power supply bus.

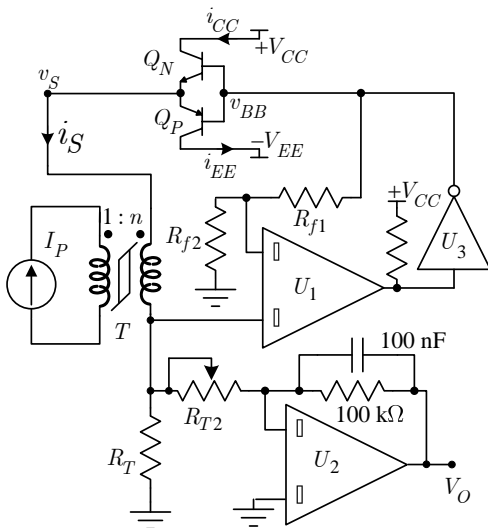


Fig 9. Self oscillating single feedback IT realization:
 $U_1 = \text{LM211}$, $U_2 = \text{TLC271}$, $U_3 = 74\text{HC1G04}$

Trimmer R_{T2} is adjusted in order to achieve overall sensitivity $V_O / I_P = 100 \text{ mV/A}$.

Since voltage v_S can have only two discrete states, naturally implied improvement for the sake of power efficiency is replacement of push-pull class B output stage with switching source-sink counterpart, Fig. 10. Although DCT in Fig 10, doesn't contain Class D CA, switching excitation of secondary winding of the flux concentrator is the same, and the transducer is prone to the same bus pumping effect.

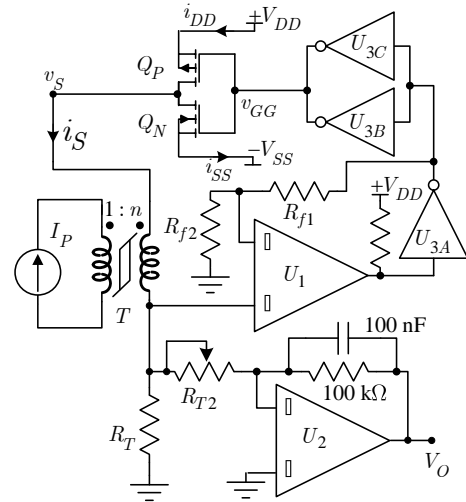


Fig. 10. Self oscillating single feedback IT realization:
 $U_1 = \text{LM211}$, $U_2 = \text{TLC271}$, $U_3 = 74\text{V2G04}$

IX. EFFECT OF THE POWER SUPPLY BUS CLAMPING

In order to prevent negative repercussions of the bus-pumping effect, some sort of artificial absorption of negative supply currents should be applied. The simplest solution is application of appropriate Zener diode as a voltage limiter, with a Zener breakdown voltage slightly higher than the maximum supply voltage, or a Zener referenced current sink [17][17]. Such solutions are supply dependent, and needs recalculation of parameters for each particularly applied power supply level.

Since the role of the clamp circuit is to absorb negative I_{D1} , (and I_{D2}) it maintains power supply buss on secure level, approximately $\approx V_{DD}$.

If $-I_{D1} < I_0$ clamping circuitry is inactive and positive portion of power supply normally delivers energy to DCT. Then, total power dissipation is given by (4), same as in ideal case. On the other hand, when $-I_{D1} \geq I_0$, positive supply is blocked, ie. $I_{DD} = 0$, and there is no energy recuperation in the power supply. The clamping circuitry is active and V_{DD} is maintained on nominal level $\approx V_{DD}$. Since only negative portion of power supply is active, the modified total dissipation P_{tot1} depends only on V_{DD} :

$$P_{tot1} = V_{DD} I_2 + V_{DD} I_{D2} \quad (16)$$

On Fig. 11, calculated dissipation is presented. Figure illustrates the case for ideal supply and the case for unilateral supply where V_{DD} is restricted by additional protection circuitry.

X. CONCLUSION

The previous analysis highlights the main drawbacks of the CA design concept in half-bridge configuration [8]-[10].

The main intention in low power design is to reduce dissipation wherever is possible, without significant functionality deterioration. One of the possibilities is reduction of P_0 by proper low power design technique to some smaller value. However, reverse current flow can be handled only by large I_1 and I_2 , since unidirectional power supply cannot help. There is low limit of I_1 and I_2 which guarantee that reversal

currents can be absorbed. One possibility is to reduce P_0 consumption in a steady state, using the appropriate technique. On the other hand, there must be enough large currents I_1 and I_2 to absorb the negative power supply, because the power supply may not be able to absorb them. In the least favorable case, there is a low limit for I_1 and I_2 , which guarantees that these currents will be absorbed by negative power supply. Consequently, the design of the DCT with CA in the form of a half-bridge amplifier in class D inherently possesses opposing requirements, i.e. the SWT design concept proposed in [8]-[10] is not generally a better option than IT, when it comes to low consumption.

In a case where overall consumption is low, there is no benefit of the switching CA topology, while linear compensating circuitry can be extremely reliable, simple and noisy-free [3]. On the other hand, when complexity of specifications and transducer electronic causes unavoidable increased standby consumption, switching CA is acceptable.

Theoretically, any bidirectional power electronic DC/DC converter (like bidirectional flyback, full bridge etc.) could overcome SWT limitations, but in practice their complexity and efficiency are reasonable for much higher powers than powers found in DCTs.

Absorption of negative current of power is certainly possible to be solved by more or less complex electronics [15], but in this case the question arises of what is being obtained and what is lost, which is an analysis that goes beyond the scope of this paper.

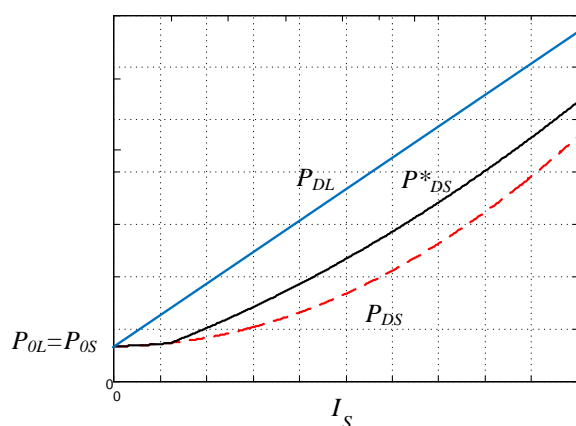


Fig. 11 Power consumption of two transducer types IT and SWT. Standby consumptions are the same, $P_{OL}=P_{OS}$, as well as other design parameters. P_{DS}^* denotes power consumption of the SWT with clamping Zener diode

When DCT is designed, principle of replacing linear CA with class D push-pull CA have two contradictory design objectives: for low power operation equivalent DC-CT conductance g_0 should be as minimal as possible, what leads to low absorption capability of negative switch's current I_{D1} or I_{D2} , whereas lowering load resistance $R_S + r_S$ in order to reduce power consumption, leads to higher maximum secondary current I_{SM} and larger bus-pumping potential.

When a DCT is applied, the user should be aware of real power demands, which are in disparity to estimates reported in [8] DCT with class D CA is significantly inferior in power efficiency compared to ideal case, laying somewhere in the middle between ideal case and the case of class AB CA.

If DC-CT is designed for low-cost commercial use, without provision of any sort of voltage limitation, or other protection mechanism against the bus-pumping, end user should be informed about appropriate protection measures. The simplest measure is specification of the minimal bleeder load which must be applied to the unilateral power supply, in order to provide bidirectional flow of the power supply currents I_{DD} and I_{SS} . The minimal load should be specified for the worst case for I_{SM} full range of allowable burden resistor R_S , and full range of allowable supply voltages. Alternatively, end user of DCT can be provided with technical information about external clamping of the power supply bus. This option offers great flexibility for the price of higher level of the end user expertise: the user should understand operation and safe operating conditions for unprotected DCT.

On the other hand, if DCT is designed for safe universal applications, some sort of internal active clamping circuitry should be provided, at the expense of increased complexity.

In both cases, comprehensive description of the power consumption, based on equations derived in the paper, should be provided to the user.

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