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Low pressure mercury lamps and half-bridge inverters

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Low pressure mercury lamps
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1. Summary

High frequency electronic ballasts are used for energy saving (more Lumens/Watt and fewer losses). The dimming range is as wide as 1-100%. At 1% dimming the electronic ballast as a power supply for fluorescent lamps has a problem. When the half bridge circuit is used, the lamp power can be regulated by changing the frequency. The discharge of the lamp however becomes unstable at low light levels and low ambient temperatures and the lamp extinguishes.

Besides the frequency dim method another method has been investigated that improve the system behaviour.

In the second method the transistor conduction time is controlled with much less instability at the low light level at low temperatures. We call it the Tt method.

The system behaviour has been simulated by using SPICE. A special lamp model has been developed. The result of the SPICE simulations are compared with the measurement results.
2. Introduction

In this report the power of fluorescent lamps will be regulated. We will consider electronic ballasts with half-bridge circuits. These ballasts regulate the power of the lamp by frequency control. The dimming of the lamp at low lamp powers is impossible without extra precautions because the lamp will extinguish [1]. In this report a method is given to regulate the power. A control loop is added to eliminate flickering of the light. In our investigation we try to find the difference between frequency control and conduction time control.

Regulating the lamp power by using a method called transistor conduction time control has disadvantages in reference to an unstable behaviour. This unstable behaviour becomes visible as irregular burning (flickering) of the lamp as it occurs to the examined lamp in a dimmed position (low levels) and at low temperatures. By controlling the half-bridge circuit in other ways and comparing these methods with each other it is possible to make a choice.

The methods which have been tested are:

[a] Frequency control
[b] Transistor conduction time control

Prediction:
A practical half-bridge electronic ballast has been made to test the electrical characteristics. It has been observed that the fluorescent lamp can be dimmed at low temperatures by Tt control with an extra feedback loop. The drive signals for the half-bridge MOS transistors may be derived from two electrical signals, which are the coil current and the pulsed voltage across one of the MOS transistors, in order to obtain zero voltage switching (low switching losses of the MOS transistors).
3. The half-bridge inverter

The static and dynamic properties of a half-bridge circuit in combination with a lamp load are difficult to analyze. Because of mathematical complexity the half-bridge circuit will be analyzed by simulation with a resistance load.

3.1 Description of the half-bridge inverter: resistive load.

In order to obtain an alternating voltage across the load, the half-bridge inverter is used. It consists of:

- Two switches: MOS transistors
- Resonant circuit: L and C
- DC blocking capacitor: C_blocking
- Load: Resistor

![fig. 3.1 Half-bridge circuit when the load is a resistor](image)
The half-bridge inverter of fig. 3.1 generates a pulsed voltage by switching both MOS transistors alternately. The situation for this investigation will be a symmetrical pulsed voltage (duty cycle of 50%). The current will be stabilized by the inductance \( L \). The blocking capacitor \( C_{\text{blocking}} \) will not allow dc voltages across the resistor. At frequencies above resonance \( (\omega_0 = 1/\sqrt{L \cdot C}) \) the voltage across the resistor can be very low. \( L \) and \( C \) are chosen in such a way that the fluorescent lamp (PL 24W) burns at nominal power (24W) when the frequency is 20 kHz. It's possible to choose \( L \) and \( C \) with the aid of the method (complex algebra) which is mentioned in the report VLc.L-VSA 9/81 [2].

### 3.2 Description of the half-bridge inverter; lamp load

The resistor in fig. 3.1 will be replaced by a fluorescent lamp. The energy will be supplied by a square wave voltage source by means of switching the MOS transistors of fig. 3.2 alternately. The lamp current will be stabilized by the inductance \( L \). In order to ignite the lamp the ignition voltage will be generated by the inductance \( L \) and the capacitance \( C \) by driving the load at a frequency near resonance (resonant rise of the voltage across the capacitor \( C \)).

![fig. 3.2 Half-bridge inverter with a lamp.](image-url)
The situation for this investigation will be a symmetrical pulsed voltage (duty cycle of 50%). In contrast with the resistance load the lamp voltage will be high at low powers. By changing the period (frequency) of the pulsed voltage the power of the lamp will be varied. When the period is decreasing, the lamp current will decrease. The voltage across the lamp will increase and the power will decrease. Under normal condition (1/ωC >> Rlamp) the lamp current is primarily determined by ωL.

3.3 Zero voltage switching of the MOS transistors

High-frequency half-bridge circuits have been commonly used because of their high efficiency. Transistors will be switched on and off in order to generate a square wave voltage. When the transistor has to be switched on or off the voltage across the concerning transistor has to be as low as possible in order to decrease the power loss denoted by:

\[
P_{\text{loss}} = \frac{1}{T} \int_{t}^{t+T} U_{ds}(t) I_{ds}(t) \, dt
\]

Zero voltage switching means \( U_{ds} = 0 \) at switching on and at switching off [8], [9], [10].

The circuit from which fig. 3.4a and fig. 3.4b have been derived consist of \( L = 2 \) mH, \( C = 10 \) nF. For both cases the same drive signals for both MOS transistors (IRF420) have been applied. The lamp however has been represented by a resistor of \( 1 \) kΩ and \( 100 \) Ω respectively. Operating the circuit at the same frequency when the lamp impedance changes (fig. 3.4a) may lead to high switching losses of the transistors. This will occur when the lamp will be ignited or dimmed. It is obvious to protect the MOS transistors against capacitive loads. As will be seen in section 6.1.1 and 6.2.1 capacitive mode includes a zero conduction time. It is also possible to determine the drive signals for the MOS transistors by a method called "zero voltage switching". With the aid of this method it is possible to allow inductive choke currents.
fig. 3.3 Half-bridge inverter: zero voltage switching of the transistors

fig. 3.4a Inverter with a capacitive load

The pulsed voltage lags the choke current.
The pulsed voltage has been generated by the inverter of fig. 3.1.

(Calculated by SPICE)
Capacitive mode

The MOS transistor however is switched (on or off) when the voltage across the transistor equals the supply voltage \( (U_{ds} = U_{supply}) \) (fig3.4a).

- When Q1 of fig. 3.3 conducts the current \( I_{\text{choke}} \) will be positive. When the current becomes negative the voltage \( U_{ds1} \) across Q1 will be the diode voltage (D1 is conducting).
- When diode D1 is conducting transistor Q2 will be switched on. The voltage across Q2 \( (U_{ds2}) \) has the same value as the supply voltage. D1 and Q2 are conducting simultaneously which results in short-circuiting the dc supply. By reverse recovery diode D1 will be switched off.
- This situation occurs too when diode D2 is conducting and transistor Q1 is switched on.
- The peak transistor currents are caused by bad timing of the drive voltages.

![Graph](image)

fig. 3.4b Inverter with an inductive load
Inductive mode
A commonly used method to drive the transistors is accomplished by adjusting the half-bridge in the inductive mode. A capacitor (Czvs) has been applied to keep the voltage across the transistor low during switching (on or off).

- When transistor Q1 conducts, the choke current will be positive. The voltage across capacitor Czvs equals the supply voltage.
- The choke current will discharge the capacitor (Czvs) slowly when Q1 is switched off. The voltage Uds1 will remain low and the transistor's current will decrease as the current through the capacitor will increase (I_choke=I_Q1 + I_Czvs). The transistor will be switched off by zero voltage switching.
- The voltage across the capacitor Czvs (U hb) reaches the zero level and diode D2 starts conducting. The choke current remains positive. Transistor Q2 can be switched on without any losses. This transistor will be switched on by zero voltage switching.
- When the choke current becomes negative transistor Q2 is already conducting.

4. The low-pressure mercury lamp

Dimming of the fluorescent lamp is more difficult to accomplish than this is the case by dimming a resistance that remains constant as a function of dissipated power. There is a non linear dependence between L_lamp rms and V_lamp rms in addition to P_lamp=P_lamp(f); it is not an unmistakable relation but a multiple value function. The ambient temperature of the lamp has a main importance in this matter. At low dim levels (P_lamp < 7W) and low temperatures (T < 10 °C) some difficulties arise by a non stable light output (flickering). Because it is desirable to dim the low-pressure mercury lamp at a very low power it is necessary to make it a particular study.

As we will find in sections 5., 6. and 7. the several transfer curves ( P=P(f), P+P(Tt) ) differ if we replace the resistor by a lamp. We can devide the difference between lamp and resistor in the next cases:

1- The behaviour of the lamp considered within a high frequency period of the lamp current.
2- The lamp considered as a resistor to which R_lamp = V_lamp rms/L_lamp rms is the resistance we adjust if the effective value ( ≈ rms value ) is calculated within one period of the lamp current.
4.1 The behaviour of the lamp considered within one high
frequency period.

It is useful to describe the lamp behaviour within a high frequency period (instantaneous values of lamp voltage and lamp current) if the behaviour of the lamp (considered within this period) changes the switching times of the MOS transistors. We can replace the lamp by a resistor if the electrical behaviour of the lamp doesn't change properties like transistor conduction time and diode conduction time. The lamp is replaced by a resistor which is calculated from \( R_{\text{lamp}} = \frac{V_{\text{lamp}} \text{rms}}{I_{\text{lamp}} \text{rms}} \). To observe possible differences between lamp and resistor, lamp voltage and lamp current have been measured in a half-bridge setup. From measurements and comparable calculations it is proved that the lamp can be replaced by a resistor.

---

fig. 4.1 An electrical equivalent for the lamp circuit
in order to determine high frequency switching behaviour
Fig. 4.2 Non linear properties of the PL 24W lamp; P=20W

The coil of the circuit in fig. 3.2 is replaced by its high frequency equivalent (fig. 4.1). The equivalent consists of $L_1=2.8 \text{ mH}$, $C_1=10.3 \text{ pF}$ and $R_1=1.4 \Omega$ and they are mentioned in the circuit of fig. 4.1. When the power is varied the resistance of the lamp (slope of the $V_{\text{lamm}}=V_{\text{lamm}}(I_{\text{lamm}})$) changes.

Fig. 4.3 Conduction times of the lamp load (calculated by SPICE).

Difference between
- conduction times : negiglible

Difference in :
- power : 0.2 W
The difference between the conduction times of the transistor and the diode is not demonstrable when the lamp is replaced by a resistor. The difference between the measured and the calculated power is 0.2W when the lamp is adjusted to 20W. The differences between the practical and the measured conduction time are negligible.

4.2 The behaviour of the lamp expressed by rms values.

The measured transfer between lamp current and lamp voltage or lamp power is represented in fig. 4.4.

![Graph](figure)

fig. 4.4a $V_{\text{lamp}} = V_{\text{lamp}}(I_{\text{lamp}})$
It is clearly to be seen in fig. 4.4a that the transfer between lamp voltage and lamp current isn't linear but has an exponential behaviour. It attracts attention that the lamp voltage has a maximum value. At low temperatures however the lamp voltage can increase considerably. The disadvantages of high voltages are mentioned in section 5.

Some mathematical models for the fluorescent lamp are treated in the reports of Vos, Reijnaerts and Morgan [3],[4], [5], [6].

The first characteristic connection between electrical properties is the \( V_{\text{lamp}} = V_{\text{lamp}}(I_{\text{lamp}}) \) relationship (Reijnaerts). This function is not able to cross the origin but has high voltages at low currents. At low currents the lamp has a high impedance \( R_{\text{lamp}} = V_{\text{lamp, rms}} / I_{\text{lamp, rms}} \). At very low currents this impedance will decrease. The gradient of the curve will be higher when the temperature of the lamp decreases. The \( V_{\text{lamp}} = V_{\text{lamp}}(I_{\text{lamp}}) \) graph may be estimated by means of the following functions:

\[
V_{\text{lamp}} = V_o + V_1 \cdot \exp \left( \frac{(I_0-I_{\text{lamp}})}{I_1} \right)
\]

The terms \( V_o \), \( V_1 \), \( I_0 \), \( I_1 \) are temperature dependent and it is also possible to cover these terms in a new function. These functions may be expressed by a polynomial or an exponential power:

\[
V_o = V_{oo} + V_{o1} \cdot T + V_{o2} \cdot T^2 + V_{o3} \cdot T^3 + \ldots
\]

or

\[
V_o = V_{oo} + V_{o1} \cdot \exp \left( \frac{T_o(T)}{T_1} \right)
\]
It has been found effective to enlarge the used methods of Reijnaerts. It has also been found useful to calculate the electrical quantities (lamp current and lamp voltage) for use in Spice. The transfer between lamp voltage and lamp current for a PL 24W lamp belonging to one temperature are as follows:

\[
V = V(I) = \Phi_0 + V_1 \exp\left(\frac{-(I-I_0)}{I_1}\right) + V_2 \exp\left(\frac{-(I-I_0)}{I_2}\right)
\]

\[
I = I(V) = \Psi_0 + I_1 \exp\left(\frac{-(V-V_0)}{V_1}\right) + I_2 \exp\left(\frac{-(V-V_0)}{V_2}\right)
\]

The parameters mentioned in these formulas depend on temperature. For one temperature, the values for a PL 24W lamp are:

\[
V(-10^\circ C) = 34.76 + 108.4 \exp\left(-\frac{1}{15.84 \times 10^{-3}}\right) + 95.55 \exp\left(-\frac{1}{137.8 \times 10^{-3}}\right)
\]

\[
I(-10^\circ C) = 0.664 \exp\left(-\frac{(V-34.1)}{16.61}\right) + 0.031 \exp\left(-\frac{(V-34.1)}{37.3}\right)
\]

![fig. 4.5 Electrical properties of a PL 24W lamp](image)

The properties of a fluorescent lamp can be expressed in the way of fig. 4.5. The measurements of the PL 24W are corresponding the method used in the report "Lamps I 4007/92" [7]. When the lamp has to be dimmed a certain curve has to be traced on the surface. It is possible with this surface to get an impression about the temperature dependence of the lamp. At certain values of the temperature the voltage has a maximum. Unfortunately it was not possible to measure the behaviour of the lamp at low light levels and low temperatures because frequency control was used (the Tt control was in development).
By increasing the power from a dimmed situation to nominal power another curve on the surface will be taken. At nominal power the lamp temperature (also called cold spot temperature) will be high. When a new, dimmed, situation has to be reached the thermal properties of the lamp in combination with a luminare will influence the electrical properties. After several (≈ 10) minutes the final state will be reached. A disadvantage of the temperature fluctuations is the light output shift when a lamp reaches its final temperature.

The time constant of the PL lamp may be taken into account. It has to be verified for different applications where a certain regulating characteristic has to be achieved in order to solve this problem.

In order to measure the time constant of the lamp in a practical situation the PL lamp has been mounted in a luminare. For ± 6 minutes the temperature of the lamp has been measured when the lamp power is changed by a step function (0.5W to 18W and 18W to 0.5W respectively).

Figure 4.6a shows the time dependency of the lamp/luminare combination. Lamp power has been measured by a Tektronix 2430a oscilloscope. Every 2 seconds the data generated by the oscilloscope has been read out by a computer (GPIB interface bus).

![Figure 4.6a](image_url)
Lamp properties influence the circuit behaviour, especially when low ambient temperatures of the lamp appear. When the frequency controls the lamp power there is no unambiguous power consumption of the lamp circuit possible at low temperatures and at low lamp powers; the lamp flickers so that another way of controlling the lamp power is necessary. When the lamp properties change by temperature variation the drive signals for the transistors have to be adjusted in order to obtain the same power. It is necessary to add the lamp time constants to determine the dynamic behaviour of the half-bridge circuit.

5. Power control by frequency variation

5.1 Resistive load

By regulating the frequency (F control) the transistor conduction time and the diode conduction time depend separately on the inverter's load. Varying the impedance of the load will change the transistor time and the diode time but the period (or frequency) of the pulsed voltage and choke current of fig. 3.1 will remain constant.

The half-bridge inverter of fig. 3.1 has been loaded with a resistor. For one combination of L and C the electrical properties are calculated to obtain nominal power at the nominal frequency. With different resistor values the power has been varied. The power is a function of the frequency and the resistances.
To explain frequency variation we take a voltage which increases linearly as a function of time (fig. 5.3). The sawtooth obtained in this way is used to determine the moment that a transistor has to be switched off. The dim level determines the frequency. Figure 5.3 shows when the transistors have to be driven. Both transistors have to be switched in the same way.

A transistor of fig. 3.1 has to be switched on within the hatched part of the drive signal (Q1 or Q2) denoted in fig. 5.2. During the time the internal body diode conducts, the voltage across the MOS transistor Uds (drain-source voltage) will be equal to the diode voltage. Switching on losses will be minimal when the transistor is switched in this way. If the transistor is switched on later than in the denoted hatched part then the voltage across the transistor (Uds) will increase and switch on losses will occur (switch off losses of the diode will increase too). These losses must be avoided because the efficiency of the half-bridge inverter will decrease by this. The power towards the load of the half-bridge inverter can be regulated with the aid of frequency control. As will be observed frequency control is not able to regulate the lamp power at low light levels and low temperatures.

![Graph showing drive signals for the half-bridge inverter](image_url)

fig. 5.3 Determining the drive signals for the half-bridge inverter
fig 5.4a Transfer of the inverter with a resistive load (high resistor values)

fig 5.4b Transfer of the inverter with a resistive load (low resistor values)

The slope of the $P=P(f)$ transfer can be both positive and negative.
Figures 5.4a and 5.4b show $P=P(f)$. It is clearly to be seen that for every resistance value the transfer from frequency to power has no ambiguities. For every frequency only one power value occurs.

If a resistance value is chosen which is about the same as $1\text{k}\Omega$ then it will attract attention that the slope of this transfer curve will change from positive to negative. If we choose another LC-combination this phenomenon will occur at another frequency. The chosen resonant circuit of which the transfer has been given in fig. 5.4a and fig. 5.4b consists of $L=2.8\text{ mH}$, $C=3.3\text{ nF}$ and the chosen supply voltage has an amount of $300\text{V}$. The transfer has been calculated from $20\text{kHz}$.

Several curves have been interrupted and will not continue till the mentioned $20\text{kHz}$. For the curves in fig. 5.4a it has been observed that a forbidden range exist when frequencies lower than about $55\text{kHz}$ are applied. Critical damping of the RLC tank circuit is obtained by the formula $R=1/2\sqrt{L/C}$ (formula A3.1). For resistor values lower than $460\ \Omega$ the choke current has an exponential shape. For resistances above $460\ \Omega$ resonance will occur (see section A3) and capacitive and inductive modes appear.

The slope of the curves (resistances above $2\text{k}\Omega$) can be very steep. In this case the power has to be regulated over a small frequency interval. At low resistances we see that the frequency interval is large when the power has to be regulated from $100\%$ to $1\%$.

A first impression we have obtained is the possibility of power regulation when a resistor has been used. For low resistance values a large frequency interval is needed and for high resistances a small frequency interval is needed to regulate the lamp power.

### 5.2 Frequency variation; lamp load

The lamp properties are normally expressed by $U_{\text{lamp}}$ and $I_{\text{lamp}}$. By calculating:

$$P_{\text{lamp}} = \frac{1}{T} \int u_{\text{lamp}}(t) * i_{\text{lamp}}(t) \, dt$$  \hspace{1cm} (obtained from fig. 4.4a)

and

$$R_{\text{lamp}} = \frac{U_{\text{lamp}}}{I_{\text{lamp}}}$$

the characteristical properties of the lamp can be added to the $P_{\text{lamp}}=P_{\text{lamp}}(f)$ graph which is shown in fig. 5.7. It is possible to replace the lamp by a resistor for every setting.
fig. 5.7 $P_{\text{lamp}}=P_{\text{lamp}}(f)$

The curves represent the electrical properties of a half-bridge inverter as mentioned in fig. 3.2 of which the load consists of a PL 24W lamp and a coil $L=2.8$ mH and a capacitor $C=3.3$ nF are used. It has been observed that for temperatures lower than $15 \, ^{\circ}\text{C}$ the dimming with the aid of frequency control is not possible.
fig. 5.8 $R_{\text{lamp}}=R_{\text{lamp}}(P_{\text{lamp}})$

fig. 5.9 $U_{\text{lamp}}=U_{\text{lamp}}(I_{\text{lamp}})$
It is not possible to obtain a lower current than ±55 mA when frequency control is used. The lamp impedance has been calculated with the formula \( R_{\text{lamp}} = \frac{U_{\text{lamp}}}{I_{\text{lamp}}} \) (rms values). In figure 5.8 the transfer \( R_{\text{lamp}} = R_{\text{lamp}}(P_{\text{lamp}}) \) is given. This figure shows the changes in the lamp impedance between ±40Ω and 20 kΩ just as is the case in fig. 5.4. The graphics are plotted on a semi-log scale to distinguish the differences at several temperatures. From this figure it is not possible to derive the extinction of the lamp at low temperatures. It is clear that the lamp impedance increases enormously at low powers. Only the curve which the \( U_{\text{lamp}} = U_{\text{lamp}}(I_{\text{lamp}}) \) at the temperature \( T = 45 \, ^\circ \text{C} \) can be dimmed to very low powers.

### 5.3 Frequency variation: Control

The behaviour of the half-bridge circuit will be explained with fig 5.10. At different values of the frequency the resistance is changed to obtain the transfer \( V_{\text{load}} = V_{\text{load}}(I_{\text{load}}) \) (from low values to high values). A measured transfer of \( V_{\text{lamp}} = V_{\text{lamp}}(I_{\text{lamp}}) \) has been added to fig. 5.10. The intersection between the resistance and the lamp transfer is a possible setting for the lamp. The lamp curve intersects two times the \( f = 75 \, \text{kHz} \) curve (multi value function). For frequencies higher than 75 kHz there is no intersection with the lamp transfer (in this case the lamp will extinguish). At 80 kHz it is not possible to ignite the lamp even if the resistance becomes infinite. The lamp will extinguish by:

- the situation of the lamp curve (different temperatures)
- the lamp voltage

![Graph of V_load vs. I_load for different frequencies](image)

fig. 5.10 \( V_{\text{load}} = V_{\text{load}}(I_{\text{load}}) \)
6. Power control by transistor conduction time variation

When transistor conduction time control (Tt control) is used the conduction time of the transistors will control the lamp power. This time (Tt) does not include the conduction time of the diode. Under practical circumstances this diode may be the body diode (it is also possible to use a separate diode as shown in fig.3.1). For the investigation it is not necessary to use a MOS transistor, it may as well be a bipolar junction transistor with an anti parallel diode.

In order to try to solve the problem (extinguishing of the lamp) which occurs by frequency control we investigate a different parameter which is able to control the power. We call this parameter transistor conduction time (Tt). We choose a signal which increases linearly in respect with time that will be started at the moment when:

- The half-bridge voltage \((U_{hb} \text{ of fig. 3.1})\) has reached the supply voltage and the choke current \((I_{choke})\) crosses zero.
- The half-bridge voltage \((U_{hb} \text{ of fig. 3.1})\) has become zero and the choke current \((I_{choke})\) crosses zero.

![Diagram](image)

fig. 6.1 Determining the drive signals for the half-bridge inverter with Tt control

The investigated half-bridge circuit consisting of \(L=2.8 \text{ mH}, C=3.3 \text{ nF}\) and a resistive load. The transistor conduction time has been varied. In the same figure the diode conduction time is mentioned. By controlling the conduction time of the transistor we can regulate the lamp power.
The concerning transistor of fig. 3.1 has to be switched on within the hatched part of the drive signal (Q_hi or Q_lo) mentioned in fig. 6.1. If the sawtooth shaped signal, which is linearly dependent on the transistor conduction time, reaches a dc level that represents the Tt, the concerning transistor will be switched off. By adjusting the reset level at different values a different transistor conduction time (Tt) will be obtained.

In fig. 6.1 a signal to denote the conduction time of the diode (Td) has also been given. In sections 6.1.1 and 6.2.1 we will see that it is not wise to regulate the diode conduction time because of plural solutions of the transfer to power.

6.1 Resistive load

As we can see from fig. 6.2a and fig. 6.2b it is possible to regulate the power of the resistance load. The curves in fig. 6.2a have been interrupted. It has been observed that the inverter operates in the capacitive mode when the dashed line is crossed. In this area no zero voltage switching is possible. The slope of the P=P(Tt) transfer is at low Tt values positive and at high values negative. As we have seen in section 5.1 "Frequency variation", we allow a change in the sign of the slope of the transfer.

![Graph showing load power vs. transistor conduction time]  

fig. 6.2a P=P(Tt); the half-bridge is loaded with a resistor (high resistance values)
fig. 6.2b \( P = P(T_t) \); the half-bridge is loaded with a resistor (low resistance values)

The assignment mentioned in fig. 6.2 don't include any plural solutions. The transfer functions from transistor conduction time to power denote that the resistor power can be regulated. For resistance values lower than \( 1k\Omega \), for this RLC-combination, it is impossible to regulate the power by a small interval of the transistor conduction time. Regarding the occurring frequency variation the frequency interval can be mentioned large. The largest value of the transistor conduction time at any resistance curve has been chosen in such way that a frequency lower than 20kHz cannot occur.

### 6.1.1 Resistive load and diode conduction time

According [1] it has been shown to be possible to calculate the power within 20% accuracy with the aid of complex algebra (this method is only valid for sinusoidal signals). For our purpose we have not chosen for this method but we want to determine the power of the half-bridge inverter by simulation (SPICE) and by formula.

To calculate the power by formula we want to make a comparison between the Spice method and the calculation. The power has been calculated when a square wave voltage is used and the choke current is assumed sinusoidal. The load consists of a resistor (fig. 6.3).
The power of the in fig. 6.3 mentioned circuit can be settled by:

\[ P = \frac{1}{T} \int_{\tau}^{\tau+T} U_{\text{pulse}}(t) I_{\text{choke}}(t) \, dt \]

\[ P = 2U_0 I_0 \cos(\omega T_d) \quad 0 \leq T_d \leq \pi/(2\omega) \quad \text{formula 6.1.1} \]

By using transistor conduction time control the diode conduction time can be measured. Fig. 6.4 shows the $T_d = T_d(P)$ relationship.
fig. 6.4a $T_d = T_d(P)$; high resistance values

fig. 6.4b $T_d = T_d(P)$; low resistance values
In fig. 6.4 the transfer $T_d = T_d(P)$ is shown when the voltage source has a pulsed shape (this is also the case when a half-bridge inverter is used). The transfer between diode conduction time control and power is measured by putting the lamp circuit in a special state by $T_t$ control. It is clearly visible that at a possible adjustment of only one value of the power more than one value can occur in the diode conduction time. The transfer from power to conduction time is a multi value function. It is not usefull to adjust the power of a resistance with the aid of diode conduction time. When we compare formula 6.1.1 with figure 6.4 than the $\omega T_d = \omega T_d(P)$ assignment doesn't look like a cosine. It is not possible to use complex algebra to determine the conduction time of the diode in all cases.

6.2 Lamp load

It is possible to use the transistor conduction time control ($T_t$ control) in order to dim the low pressure discharge lamp at low luminance levels and low temperatures. By decreasing the transistor conduction time the power will also decrease. The $T_t = T_t(P)$ transfer makes it possible to adjust the power with the same slope. However the transfer is not linear. Below 25% of the nominal power the slope of the transfer between the transistor conduction time and the power is very steep. In this way dimming of the lamp at low powers is not impossible but remains difficult. The variation of the transistor conduction time is very small when the power has to be regulated from 25% of the nominal power to 1% (the lamp gives only a little light). In fig 6.5 the $P = P(T_t)$ transfer is given. Therefore a special designed feedback loop has been added.

fig. 6.5 $P = P(T_t)$; lamp load
6.2.1 Lamp load and diode conduction time

The resistance mentioned in section 6.1.1 is replaced by a PL 24W compact fluorescent lamp. The power has been adjusted by transistor conduction time control. At the same measurement the diode conduction time has been noticed. Figure 6.6 shows the relation between the diode conduction time (Td) and the power (P).

![Diode conduction time vs Lamp power](image)

Fig. 6.6 The Td=Td(P) assignment when the load of the inverter is a lamp

It is clearly visible that plural solutions of the power occur at one diode conduction time value. It is impossible to control the power by the diode conduction time because the transistors must be switched at high drain to source voltages (no zero voltage switching).

6.3 Control

At low temperatures and low powers it has been observed that it is impossible to dim the lamp without an extra control loop. Lamp fluctuations at low lamp currents occur because of the exponential behaviour of the \( V_{lamp} = V_{lamp}(I_{lamp}) \) transfer. The slope of the \( T_t \) transfer (load line) at short conduction times has almost the same slope of the lamp curve. It is difficult to obtain a stable operating point. An accurate setting of the transistor conduction time will be necessary. A low pass filter is used to eliminate power variations. The investigated lamp circuit (actuator) possesses not only this compensation method but an integrator too. With the designed circuit however it was not possible to make the same actuator adjustments at nominal power (24W) and in a dimmed position (1W). Figure 6.7 shows the used control loop.
Because of the steepness of the curve the power can change from 0.2 watt till 6 Watt (fig. 6.5) when the transistor conduction time changes from 3.93 µs to 3.97 µs at -10 °C. Therefore choke current Ichoke and U_{hb} of fig. 3.2 are measured to determine the transistor conduction time. Now the moding or lamp current oscillation will stop.

**fig. 6.7 Control loop to eliminate the lamp current oscillations**

**fig. 6.8 U_{load}=U_{load}(I_{Load})**

From fig. 6.8 we know that Tt control can be used to regulate the power. The adjustments were settled by the intersection of the resistance and the lamp curve. There are no double values of the power when Tt control is used.
The half-bridge circuit has been loaded with a lamp (fig. 6.9). In order to determine the operating frequency the circuit is controlled by Tt control (the transistor conduction time is used to drive the transistors of the half-bridge circuit). Now the $F=F(P)$ can be measured. Two values of the power are possible at one frequency. Another method than frequency control has to be chosen to dim the lamp at low light levels and low temperatures. It is not possible to solve this problem by only using a low pass filter in the feedback loop.
7. simulation of the half-bridge circuit (with lamp)

7.1 A model for the fluorescent lamp

As is explained in the previous sections it is possible to replace the lamp by a resistor (it is not necessary to describe the lamp within a high frequency period) and it is evident that a mathematical description for the behaviour of the lamp, by means of rms values of lamp voltage and lamp current, is sufficient. The lamp model is explained with the aid of elementary mathematical functions. The lamp is represented in the model by a resistor. The resistance is calculated from \( R_{\text{Lamp}} = V_{\text{Lamp}_{\text{rms}}} / I_{\text{Lamp}_{\text{rms}}} \). The value of the lamp voltage is calculated by means of a meter which calculates the rms value per half frequency period. Lamp current will be calculated with the aid of the transfer \( I_{\text{Lamp}_{\text{rms}}}=I_{\text{Lamp}_{\text{rms}}}(V_{\text{Lamp}_{\text{rms}}}) \). It has been proved possible to investigate the instability which occurs by dimming of fluorescent lamps.

The rms meter

According to the formulae

\[
V_{\text{rms}} = \sqrt{\frac{1}{T} \int_{0}^{T} V^2(t) \, dt}
\]

we form a model for the rms meter. In fig. 7.1 different signals are mentioned which will be needed for the explanation.

---

The input voltage \( U_{\text{input}} \) has been squared. A pulse will be generated at every zero crossing of the input voltage.

---

fig. 7.1a Determining the signals to obtain the root mean square (rms) function
fig. 7.1b Determining the signals to obtain the root mean square (rms) function

In fig. 7.1a the squared value of the input voltage and the reset pulse are shown. At the zero crossing of the input voltage the mentioned signals of fig 7.1b (V_int and T_int) will be reset after every half period of the high frequency cycle. The top values of the signals V_int and T_int at the end of each half cycle will be used to calculate \( V_{rms}^2 \).

It has been proved from the simulation with Spice that attention must be paid to the square root function. By choosing a wrong method to accomplish this function, the condition number \([11]\) can become rather high. The condition number expresses the relative change of the output as a result of a (small) relative change by the input. If the condition number is small \((\approx 1)\) than the problem is known well conditioned; a small change at the input does not change the output dramatically.

The lamp model

The lamp model is constructed according the next formulae:

\[
I = I(V) = I_0 + I_1 \exp\left(\frac{- (V-V_0)}{V_1}\right) + I_2 \exp\left(\frac{- (V-V_0)}{V_2}\right)
\]

The transfer for \(-10^\circ C\) is expressed as:

\[
I(-10^\circ C) = 0.664 \exp\left(\frac{- (V-34.1)}{16.61}\right) + 0.031 \exp\left(\frac{- (V-34.1)}{37.3}\right)
\]

It has been proved that the exponential terms cannot be used immediately in connection with numerical instabilities; the condition number becomes too high. The condition number decreases when we replace the exponential power by the next formula:

\[
\exp\left(\frac{(V-V_0)}{V_1}\right) = A \exp\left(\frac{-(V-V_0)}{16.61}\right) = 0.94 \exp\left(\frac{-(V-V_0)}{16.61}\right)
\]

The term \((V-V_0)\) can be strongly changed because the mantissa is near the number 1.
To raise \( (V-V_o) \) in power it is not possible without using the exponential power function. That's why we choose:

\[
x^y = \exp( \ln(x^y) ) = \exp( y \ln(x) )
\]

\( \ln(x) \) : constant

\( y \) : input value

If we calculate the condition number for \( \exp( \ln(x) (V-V_o) ) = \exp(-|\ln(x) (V-V_o)|) \) than it appears that the condition number is very high. Therefore we take the inverse of \( \exp( +|\ln(x) (V-V_o)| ) \):

\[
x^y = ( \exp(-y \ln(x)) )^{-1}
\]

The model for the transfer between lamp voltage and lamp current looks as follows:

![fig. 7.2a A model for the \( Z=Z(V_{rms}) \) transfer at \( T=-10 \, ^\circ C \)](image)

![fig. 7.2b A model for the PI 24W lamp](image)

The transfer of fig. 7.2 is in good agreement with the measured values of lamp voltage and lamp current.
7.2 Power control by frequency

At the simulation the MOS transistors of fig. 3.2 are replaced by a symmetrical square wave voltage source. The period of the square wave voltage remains constant during the simulation. During the simulation the lamp keeps one temperature. In fig. 7.4 the P=P(f) transfer is given. It is clearly to be seen that some curves which correspond to low temperatures are interrupted. Simulations and measurements are in good agreement with each other.

![Graph showing P=P(f) transfer](image)

**fig. 7.4 P=P(f)**

7.3 Power control by transistor conduction time

The transistor conduction time will be kept constantly at these calculations. In this case the half-bridge circuit generates a symmetrical square wave voltage too. If the time constants which occur by the changing of the lamp impedance are kept high then we remark that the transfers P=P(Tt) are according the measured curves of fig. 6.5. Per half period the square wave is adapted to maintain the transistor conduction time.
8. Simulated properties compared with measurements

The properties of a half-bridge circuit (fig. 3.2) are investigated on reliability in reference to measurements. It appears that the results concluded from the calculations don't differ concerning the course of the curve. The difference between the measured and the calculated frequency at low powers of the $P=P(f)$ transfer is ± 3 kHz. The difference is ±0.2 μs at low powers (1W) in case of Tt control and it's ±0.1 μs in case of Tt-Td control. At the measurements transistor losses and coil losses cannot be neglected. Especially at low powers these losses can be of great importance. The losses of the transistor are estimated at 0.5 W and the losses of the coil are estimated at 0.4 W. The lamp electrodes are heated during the measurements by a separate voltage source. These losses don't influence the measurements.
9. Remarks

Some methods to adjust the power

The power of the lamp can be measured in the following manners:

-1- \( P_{\text{lamp}} = \frac{1}{T} \int_{t} u_{\text{la}(t)} i_{\text{la}(t)} \, dt \)

-2- \( P_{\text{half-bridge}} = V_{\text{supply}} \times I_{\text{osupply}} \)

The power can be simply calculated with the help of the supply voltage and the current delivered by the supply source. This is a very easy method. A measurement of the supply voltage and the dc. component of the supply current is necessary. If the supply voltage is assumed constant then it is only necessary to measure the dc. component of the supply current. This method in order to determine the lamp power is not as accurate as method 1 because transistor, coil losses and electrode loses are added.
10. Conclusion

The following adjustments are investigated

- frequency control
- Tt control

When frequency control is used it has been approved possible that under special conditions (low light levels and low temperatures) problems can arise in relation to the dimming of the low pressure mercury lamp. It is possible to make an arrangement to dim low-pressure mercury lamps with the aid of Tt. For this method it's absolutely necessary to add a feedback loop in order to eliminate light flicker. It is possible to regulate the lamp current till only a few milli amperes. It appears that the square wave voltage and the choke current of the half-bridge circuit can be used to derive the transistor drive signals.
11. Literature

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High frequency quasi-resonant and multi-resonant converter topologies

Design-oriented analysis of boost zero-voltage-switching resonant dc-dc converter

Generalized topologies of zero-voltage-switching and zero-current-switching resonant dc/dc converters
Proceedings National Aerospace and Electronics Conference (NAECON '87)
1987 vol.2 p.472-478

dictaat TUE Matthey
**A1 circuit description**

**Constant supply voltage**

The mains supply voltage has been achieved by an auto transformer. After rectifying the alternating voltage the remaining ripple is fully eliminated by a capacitor. The capacitor voltage will be called the supply voltage. The supply voltage is 300V.

**Half-bridge circuit**

The half-bridge circuit consists of:

- Two switches
- Resonant circuit
- Dc blocking capacitor
- Load

: Mos transistors
: L_choke and C_lamp
: C_blocking
: Lamp PL 24W

**Drive signals for the half-bridge transistors**

The drive signals will be applied to the gates of the MOS transistors by using a level shifter IC. At the input of this level shifter IC two signals are applied which will be further called drive signals "Q_lo" and "Q_hi". Q_lo defines the switching behaviour of transistor Qlo.

**Two control signals that determine the sign of the choke current and the square wave voltage**

The square wave voltage is detected by C_hb and R_hb. The signal V_hb(H) is the amplitude quantized version of V_hb. The choke current has been determined by R_choke. This quantity will also be quantized in amplitude with the aid of a fast comparator. (μA710: IC_2). Now we get the both amplitude quantized signals V_hb(H) and I_choke(H). With the help of these two signals the transistor conduction time and the diode conduction time will be derived.

**Tt measurement**

With the aid of a combinational inclusive OR circuit (V_hb ⊕ I_choke)(H) the voltage across capacitor C1 will be regulated. During the time that this signal is active the voltage V_c1 is increasing linear with the time. With the aid of MOS transistor Q1 (BST70A) the capacitor is discharged when V_c1 has reached a certain reset level. The constant current which charges the capacitor C1 will be achieved by Opamp IC_2.

**Diode conduction time measurement**

During the conduction time of the diode, capacitor C2 will be charged. The voltage V_c2 increases linear in respect with diode conduction time. When the diode conduction time is over capacitor C2 will be discharged by MOS transistor Q_2 (BST70A).
Feed-back loop \( T_t \) control

The power can be regulated with the aid of the methods mentioned in section 8. The switch off moment of
the transistors is achieved by \( T_t \). The power measurement is slow \( (\tau > 1\text{ms}) \) comparing a high frequent
period \( (T < 50\ \mu\text{s}) \). The \( T_t \) is equal to the adjusted reset level of capacitor \( C_1 \). It's not needed to measure
this quantity.

The top level of the voltage \( V_{c2} \) will be detected by IC\(_3\) (LF356) in order to measure \( T_d \). This top level
will be kept constantly during a high frequency period.

An integrator IC\(_4\) eliminates final errors in the power. The low frequent signal formed in this way
is used to adapt the transistor time.
fig. A1.1a Half-bridge circuit
fig. A1.1b Half-bridge circuit
fig. A1.2 Tt control loop
A2 SPICE MODELS

A model for the RMS measurement:

```plaintext
.SUBCKT RMS I 32
X1 1 0 2 LIMITER
X2 2 3 DELAY {TD=2U}
X3 2 3 4 XOR
V1 4 5 0.5
X4 5 0 6 LIMITER
R1 6 0 10MEG
X5 1 7 MUL {K=1}
R2 7 8 20K
E1 9 0 8 1E3
C1 8 9 50N
M1 8 10 9 9 SWITCH1
R3 9 0 10MEG
E2 10 9 6 0 10
E3 11 0 0 9 1
R4 11 12 1K
R5 12 0 10MEG
E4 15 0 12 13 1E3
R6 0 13 10MEG
R7 13 14 1K
V2 15 16 0.5
D1 16 14 DIOD
C2 17 0 200N
R8 14 17 100K
V3 18 0 100
R9 18 19 20K
E5 20 0 19 1E3
C3 19 20 50N
M2 19 21 20 20 SWITCH1
R10 20 0 10MEG
E6 21 20 6 0 10
E7 22 0 0 20 1
R11 22 23 1K
R12 23 0 10MEG
E8 26 0 23 24 1E3
R13 24 0 10MEG
R14 24 25 1K
V4 26 27 0.5
D2 27 25 DIOD
C4 28 0 200N
R15 25 28 100K
X6 28 29 GAIN {K=1E-2}
X7 29 30 31 MUL {K=1}
E9 32 0 17 31 1E3
X8 32 32 30 MUL {K=1}
R16 32 0 10MEG
R17 17 0 200
R18 28 0 200
.MODEL SWITCH1 NMOS( IS=0 LEVEL=3 KP=7 IS=0 VTO=2.55 )
.MODEL DIOD1 (IS=1E-12 N=.0597)
.ENDS
```
A model for the exponential function:

```plaintext
.SUBCKT EXPONENT 1 2
E1 2 0 1 0 1 1.5 1.666667 0.041667 8.333E-3 1.389E-3 1.98413E-4 2.4802E-5 2.7557E-6 2.7557E-11
.ENDS
```

A model for the ln(x)/y function:

```plaintext
.SUBCKT LNEXP 1 2 3
 X1 1 2 4 MUL { K=1 }  
 X2 4 5 EXPONENT  
 X3 3 5 6 MUL { K=1 }  
 E1 3 0 7 6 1E6  
 R1 5 0 10MEG  
 R2 7 0 10MEG  
 V1 7 0 1
```

A model for the fluorescent lamp (PL 24W at -10°C)

```plaintext
.SUBCKT PLMODEL 1 2
*  INPUT OUTPUT
*  INPUT=V _LAMP OUTPUT=R_LAMP (RMS VALUES)
 X1 1 3 4 SUM2 { K1=1 K2=-1 }  
 V1 3 0 34.115  
 X2 5 4 6 LNEXP  
 V2 5 0 -6.02E-2  
 X3 7 4 8 LNEXP  
 V3 7 0 -2.68E-2  
 X4 6 8 9 SUM2 { K1=6642 K2=03105 }  
 X5 2 9 10 MUL { K=1E4 }  
 E1 2 0 1 10 100  
 R1 2 0 10MEG  
.ENDS
```
A3 Calculation of formula A3.1

The voltage across the resistor of fig. 3.1 can be calculated with the aid of the differential equation:

\[
\frac{E}{LC} = V_c'' + V_c' \frac{R}{C} + \frac{V_c}{LC}
\]

The natural solution of this equation can be calculated with:

\[
V_c'' + V_c' \frac{R}{C} + \frac{V_c}{LC} = 0
\]

or in general:

\[
V_c'' + V_c' \beta \omega_o + \frac{V_c}{\omega_o^2} \quad \omega_o^2 = \frac{1}{LC} \quad \beta = \frac{1}{2RC\omega_o}
\]

\[
(s - p1)(s - p2) = 0
\]

\[
p1,2 = \sigma \pm \omega_n \quad \sigma = -\beta \omega_o \quad \omega_n = \omega_o \sqrt{\beta^2 - 1}
\]

\[
\omega_n^2 = \frac{1}{4R^2C^2} - \omega_o^2
\]

at critical damping:

\[
\frac{1}{4R^2C^2} - \omega_o^2 = 0 \quad R = \frac{1}{2} \sqrt{L/C}
\]