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USING A BROADBAND SIGNAL BASED ON M-SEQUENCE FOR AUTOMATIC PREVENTING OF ACOUSTIC RESONANCE IN HIGH PRESSURE DISCHARGE LAMPS

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Abstract: In the article, a method for preventing acoustic resonance of high-pressure lamps has been proposed. The main idea of which is the filing on the lamp of a modulated pseudo-random signal. For the set of a "discharge lamp - prototype of the device" an analysis of the spectral characteristics of the voltage on the lamp are made. The mathematical modeling of the set of a "discharge lamp - prototype of the device" is implemented. The results of simulation, which confirms the results of theoretical research, are presented.

Keywords: *preventing of acoustic resonance; high-pressure discharge lamp; pseudo-random signal.*

1 Introduction

In order to increase the efficiency of a gas-discharge low-pressure lamp, for a long time uses electronic ballasts that operate at high frequencies [1], [2]. Their advantages are:

- ◊ high coefficient of efficiency that significantly exceeds the similar parameter of electromagnetic ballast;
- ◊ the possibility of adjusting the current of the lamp in the specified limits;
- ◊ increase of light output;
- ◊ eliminating the possibility of lamp operation emergencies.

At the use of high-frequency electronic ballasts for work with high-intensity discharge lamps (mercury and sodium vapor, metal halide, etc.), there is a problem of the occurrence of acoustic resonance in the arc tubes of these lamps [1], [2]. Acoustic resonance has been studied in many scientific papers. Acoustic resonance frequencies range from 3 kHz to 150 kHz. The reason for this phenomenon is that when the direction of current flow changes due to the redistribution of charge carriers, acoustic waves arise which cause the forced mechanical vibrations of the gas medium of the arc tube, its walls, and even the fittings of its suspension. The spectrum of the resonant frequencies of mechanical oscillations depends on the geometric dimensions of the lamp arc tube, the speed of sound in it (which, in turn, depends on the pressure), the features of the designs of the arc tube and many other parameters [3], [4], [5]. At the some frequency, the frequency of acoustic waves in an arc may coincide with the resonant frequency of mechanical oscillations in the system.

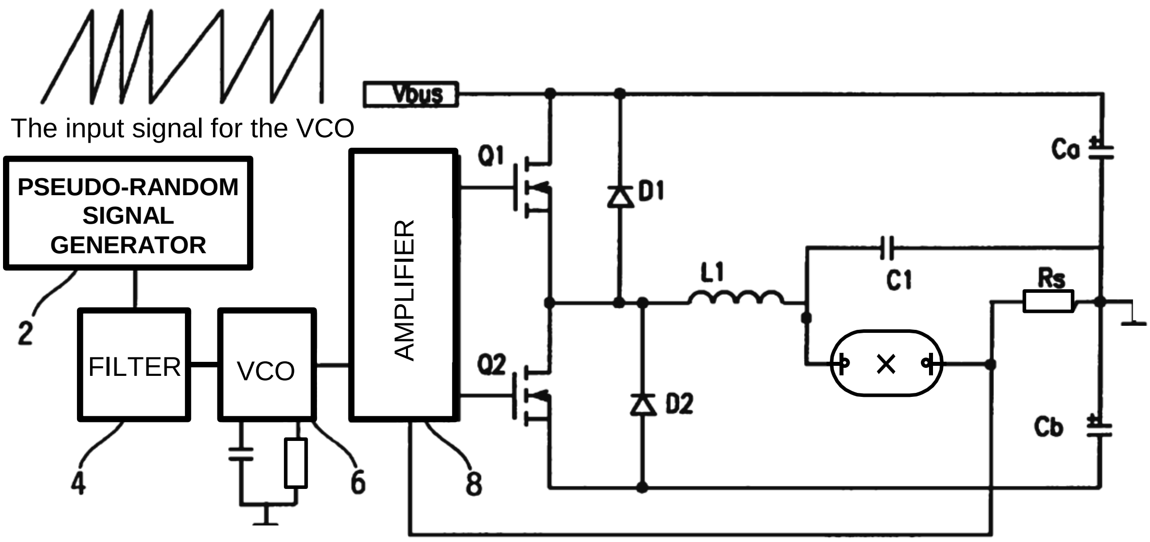

Fig. 1. Scheme of electronic ballast [6] for gas-discharge lamp:
VCO (voltage controlled oscillator)

Each of lamps, including the same type and the same power, can have different frequencies of the major resonance due to different electrical and structural parameters, which change in the direction of decrease or increase with the increase of the lamp's life time. In addition to the major resonance, there are resonances on other frequency harmonics, which negatively affects the lamp gas discharge state. The consequences of acoustic resonance are the instability of the lamp burning, the arc extinction and, at worst, the physical destruction of the arc tube [4].

Therefore, as a voltage supply of a high-pressure lamp, in most cases, alternating voltage is used, the frequency of which does not exceed 200 Hz, thus excluding the possibility of an acoustic resonance. These operating frequencies lie in the audio range and therefore the process of lamp discharges is accompanied by significant interfering noises. In addition, the size of the reactive elements used in the lamp's switching scheme remains significant compared to the ballasts at 50/60 Hz. Therefore, the search for ways of supplying high-pressure gas discharge lamps is still relevant.

The one of the effective methods for controlling acoustic resonance is the use of modulation of the main frequency of the supply voltage of the lamp by another signal. One of these schemes (US Patent No 6,144,172) [6] (Fig. 1) contains a white noise generator, which is based on a linear feedback shift register, the output signal of which flows through a filter to a voltage-controlled variable frequency generator (VCO). The signal from the generator's output enters to the output amplifier and is fed to the circuit of the inductive ballast, which are constructed in the usual way. The frequency range of the frequency of the VCO generator is usually chosen small. As a result, the output frequency of the signal coming to the lamp is changed by some pseudorandom behavior. The frequency of the change of the output signal of the pseudorandom signal generator is chosen much less than the average frequency of the VCO.

When working on the circuit, the frequency of the signal on the lamp is constantly changing and the resonance of the arc tube has no time to develop due to the inertia of the processes in the lamp.

The analysis of the described scheme allows us to conclude that in the case when a broadband signal is presented to a lamp, the energy distributed in a dangerous frequency interval will be small and the acoustic resonance phenomena will not occur or will have little effect.

The signal must also meet these requirements:

- ◊ the average value of the signal must be zero; otherwise, the phenomenon of migration of ions to the one of the electrodes of a lamp may occur;
- ◊ the duration of the constant state at the lamp inlet does not exceed the set value T, which is connected with the current limitation by means of reactive elements.

2 The offered device

The block diagram of the offered device is shown in Fig. 2. It consists of a clock generator (Gen), which generates at its output a meander at frequency of 10-40 kHz. This meander is fed to the input of the modulator sequence shaper (Mod Seq) and to the frequency divider on K (Div K). The divider generates a synchronization signal of the pseudo-random sequence generator based on the linear feedback shift register (LFSR) and simultaneously outputs to the inputs of the modulator sequence generator the clock number from the moment of switching the LFSR register.

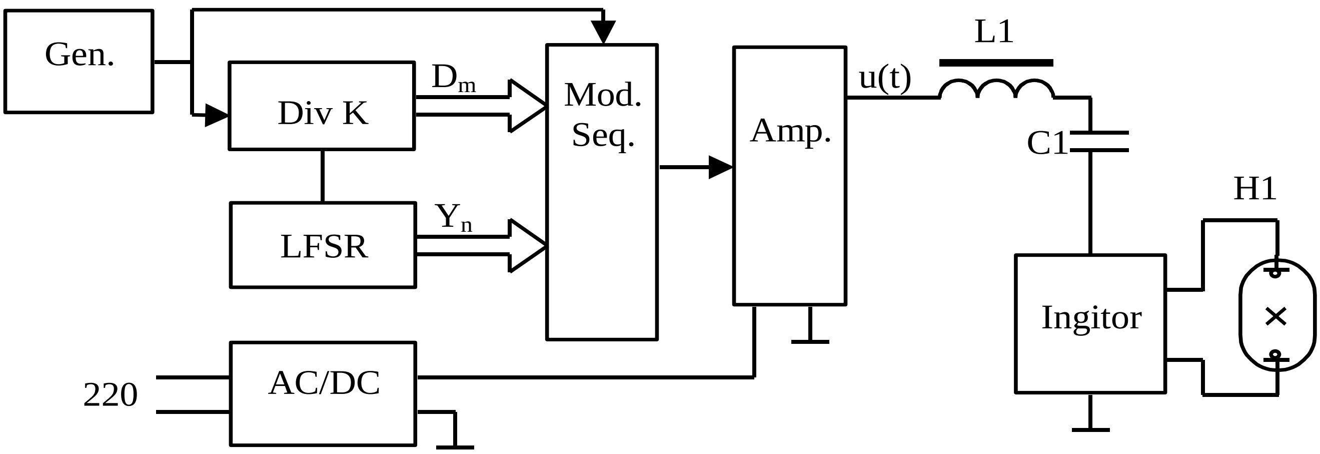

Fig. 2. Block diagram of the offered device

The modulator sequence generator (Mod Seq) also receives a signal from the LFSR register, which shows the current and previous state of the register LFSR. The modulator sequence generator based on the above signals generates an output signal to the Amp amplifier. The modulated signal at the output of the last circuit will be determined by both the current state of the LFSR register and the previous state of the register. The amplifier (Amp) operates in switch mode, the output signal of which is fed through the ballast circuit (L1 C1), and, if necessary, the ignition device (Ingitor), and the lamp H1.

Structure of random signal generator with linear feedback and signal properties

An LFSR generator (Fig. 3) consists of, the shift register 1, synchronized with pulses received from divider, and the feedback circuit 2, which calculates the value of the next bit, which enters the information inputs (A and B) of the sixteen bit register DD1/DD2. The register consists of two eight-bit registers that have a common CLK clock signal.
The number of flip-flop circuits, which are covered by feedback circuit, considered as register length and denoted as N. The bits of the cells will be numbered *i=0, 1,..., N-1*, the contents of the cell with the number and denoted as Qi. The new bit value Q_0 is determined by the shift of the bits in the register by the feedback circuit. The feedback circuit function is a linear Boolean function from the values of some case bits of register. The function executes the multiplication of the register bits by the coefficients and the "Exclusive-OR" operation above the multiplication results:

$$A = \overline{\bigoplus_{i=0}^{N-1} \overline{c_i} \vee Q_i} \ . \tag{1}$$

The number of coefficients coincides with the number of bits in the register, the coefficients c_i take the value $\{0,1\}$, with the coefficient $c_{N-1} = 1$, and the remaining coefficients are selected in a special way to obtain a given sequence length.

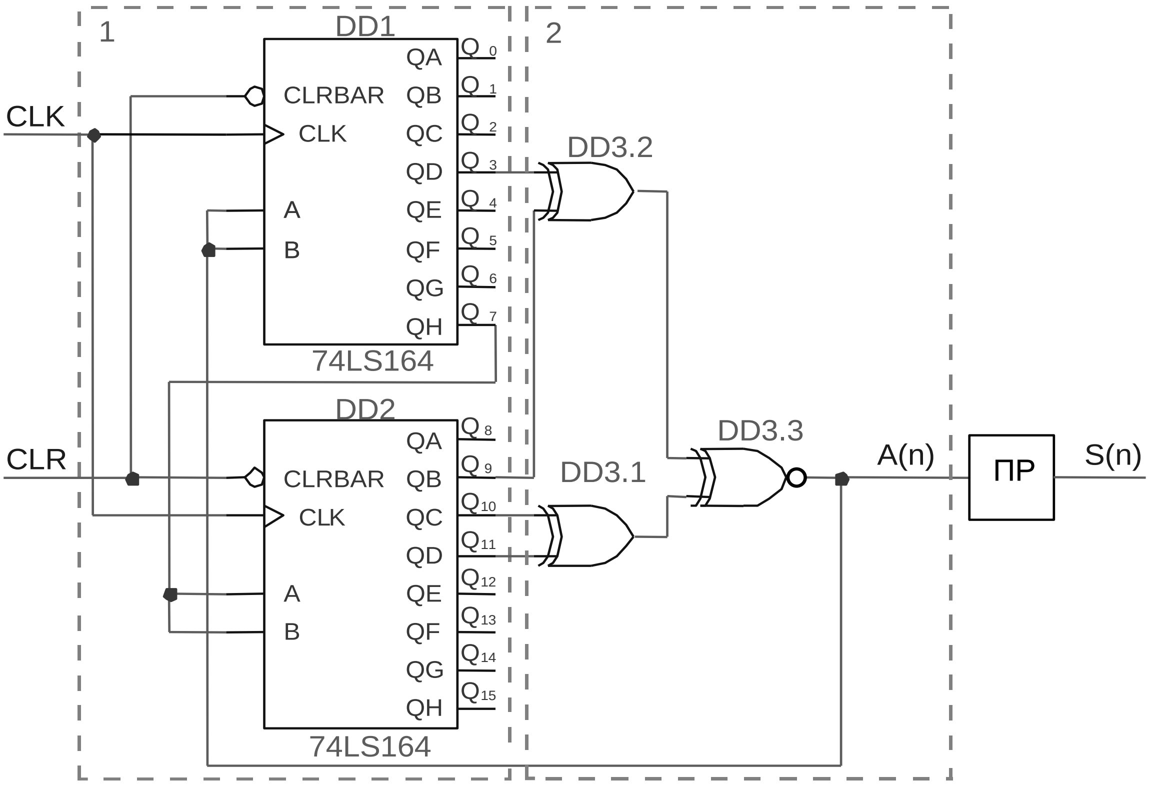

Fig. 3. Generator based on a linear feedback shift register LFSR (N = 12)

During each step, the linear feedback shift register performs the following operations:

- ◊ reads the bit placed in the $N-1$-th flip-flop. This bit is the next bit of the output sequence;
- ◊ the feedback circuit function calculates a new value for a zero flip-flop circuit using the values of other bits by the corresponding formula;
- ◊ the content of each flip-flop moves to the next cell;
- ◊ the value of the bit calculated earlier is written to into zero trigger.

Obviously, the state of "all units" at register output can generate only a single value output feedback circuit, resulting in a sequence, which consists only of the units, so getting to this state should be deleted. A shifting register has 2^N initial states, which are given by the same combination of bits in the register. Consequently, the number of permissible states is $2^N - 1$, and the maximum sequence has a period not exceeding $M = 2^N - 1$.
Consider a signal $S(k)$ that is built based on signal $A(k)$ (signal $S(k)$ takes values -1 or +1, and the signal $A(k)$ is 0 or 1, respectively):

$$S(k) = 2A(k) - 1 \, . \tag{2}$$

The sequence $S(k)$ with a maximum length $M = 2^N - 1$ at the output of the circuit is called the M-sequence [7]. For signal $S(n)$ S (n) discrete autocorrelation function:

$$K_S(d) = \lim_{L\to\infty} \frac{1}{2L} \sum_{n=-L}^{L} S(n)\overline{S(n-d)} = \begin{cases} 1 & \text{for } d = kM; \\ -1/M & \text{for } d \neq kM; \end{cases} \tag{3}$$

where k - integer number, M -sequence period. Note that the discrete autocorrelation function can be written as:

$$K_S(d) = \frac{1}{M} \sum_{n=0}^{M-1} S(n)\overline{S(n-d)}. \tag{4}$$

3 Modulator sequence generator and output signal representation

As denoted above, the signal at the output of the modulating sequence generator determined by the current state and N-1 previous states of the output of the LSFR register, and state of divider. Then the signal at the output of the modulating sequence generator is defined as:

$$u(t) = \sum_{i=0}^{P-1} S(n(t)-i)\, q\left(\frac{t}{\Delta T} - i\right), \tag{5}$$

where $n(t)$ - the LSFR register state number at the time t. $q\left(\frac{t}{\Delta T}\right)$ - some filling signal, which is equal to 0 outside the range from 0 to $\Delta T\ N$.

The filling signal must have the following properties:

1) the signal becomes -1, 0 or 1;
2) must be: $q(\lambda)q(\lambda+n) \equiv 0,$ for all λ, if integer $n \neq 0$;
3) for all $\lambda \notin [0,N]$ ($N \leq M$), $q(\lambda) = 0$.

Introduce the function $\psi(\eta)$ equal to 1 at the interval $t \in [0,1/P)$. Then the signal can be represented in the form:

$$q(t/\Delta T) = \sum_{n=0}^{P-1} A_n\, \psi\left(\frac{t}{\Delta T} - \frac{n}{P} - B_n\right), \tag{6}$$

where A_n takes values -1, 0 and 1, B_n -changes from 0 to $N-1$. Then the condition of non-overlapping of the function with its copies shifted by $\Delta T\ k$ is satisfied automatically, and condition 3 is satisfied at $N \leq M$. The signal can also be recorded as a convolution:

$$u(t) = U_0 \int_{-\infty}^{\infty} q(\tau/\Delta T) s(t-\tau)\, d\tau\ , \tag{7}$$

where $s(t,L) = \lim_{L\to\infty} \sum_{n=-L}^{L-1} S(n)\delta(t - \Delta T n)$ - representation of the impulse sequence through a set of Dirac delta functions. Indeed, substituting the impulse sequence into the equation and considering that $q(\lambda) = 0$ if $\lambda \notin [0,N]$, $|q(\lambda)| \leq 1$ and using the filtering property of delta functions, we can rewrite signal as:

$$u(t) = U_0 \sum_{n=-\infty}^{\infty} S(n) q(t/\Delta T - n) = \sum_{i=0}^{P-1} S(n(t)-i)\, q\left(\frac{t}{\Delta T} - i\right). \tag{8}$$

This expression can be written by selecting the terms of each period separately, due to the periodicity of the sequence, the expression can be written as:

$$u(t) = U_0 \sum_{n=0}^{M-1} S(n) \left(\sum_{l=-\infty}^{\infty} q(t/\Delta T - n + M\,l) \right). \tag{9}$$

Representation of the signal u (t) as a Fourier series

Since the sequence $S(n)$ is periodic with the period M, and the interval of nonzero values $q(t/\Delta T)$ is limited, it is easy to prove that the signal $u(t)$ is also periodic with the period $T_u = M\Delta T$, and can be decomposed into a Fourier series by frequencies $\omega_k = \frac{2\pi k}{\Delta T\, M}$. Using representations $u(t)$ through (9) we obtain:

$$U(\omega_k) = \frac{U_0}{T_u}\sum_{n=0}^{M-1} S(n)\int_0^{T_u}\sum_{l=-\infty}^{\infty} q(t/\Delta T - n + M\, l)\exp(-j\omega_k t)dt\,. \quad (10)$$

The internal sum and $\exp(-j\omega_k t)$ has a period equal $M\;\Delta T$, therefore, the value of the integral will not change while the integration boundaries shifted:

$$U(\omega_k) = \frac{U_0}{T_u}\sum_{n=0}^{M-1} S(n)\int_{n\,\Delta T}^{T_u + n\,\Delta T}\sum_{l=-\infty}^{\infty} q(t/\Delta T - n + M\, l)\exp(-j\omega_k t)dt\,. \quad (11)$$

Now change the order of integration and introduce the replacement of variables $t/\Delta T - n + Ml = \eta$:

$$U(\omega_k) = \frac{U_0 \Delta T}{T_u}\sum_{n=0}^{M-1} S(n)\sum_{l=-\infty}^{\infty}\int_{M\,l}^{T_u/\Delta T + M\,l} q(\eta)\exp(-j\omega_k \Delta T(\eta + n - Ml))d\eta\,. \quad (12)$$

Again, taking into account the frequency of the expression $\exp(-j\omega_k \Delta T(\eta + n)$ in the variable n with a period M and the value of $T_u = M\;\Delta T$, we get:

$$U(\omega_k) = \frac{U_0}{M}\sum_{n=0}^{M-1} S(n)\sum_{l=-\infty}^{\infty}\int_{M\,l}^{M(l+1)} q(\eta)\exp(-j\omega_k \Delta T(\eta + n))d\eta\,. \quad (13)$$

From the definition of the $q(\eta)$ it follows that under the sign of the internal sum there will be only one non-zero term, which corresponds to $l = 0$:

$$U(\omega_k) = \frac{U_0}{M}\sum_{n=0}^{M-1} S(n)\int_0^{M} q(\eta)\exp(-j\omega_k \Delta T(\eta + n))d\eta\,, \quad (14)$$

grouping the factors we get:

$$U(\omega_k) = \frac{U_0 Q(\omega_k)}{M}\sum_{n=0}^{M-1} S(n)\exp(-j\omega_k \Delta T n)\,, \quad (15)$$

where:

$$Q(\omega_k) = \int_0^{M} q(\eta)\exp(-j\omega_k \Delta T \eta)d\eta = \int_{-\infty}^{\infty} q(\eta)\exp(-j\omega_k \Delta T\,\eta)d\eta\,, \quad (16)$$

it was taken into account that the function $q(\eta)$ is equal to zero outside the interval $[0..M]$.

Fourier series representation of the signal autocorrelation function

Proceed to consider the continuous autocorrelation function of the signal, which we write for the periodic continuous signal in the following form.

$$K_u(\tau) = \frac{1}{T_u}\int_0^{T_u} u(t)\overline{u(t-\tau)}dt\,. \quad (17)$$

The signal $u(t)$ are limited, so the $K_u(\tau)$ value are also limited. Due to the periodicity $u(t)$, the autocorrelation function is also periodic with period T_u, and can be decomposed into a complex Fourier series at frequencies $\omega_k = \frac{2\pi k}{T_u} = \frac{2\pi k}{M\Delta T}$ with coefficients:

$$K_u(\omega_k) = \frac{1}{T_u}\int_0^{T_u} K_u(\tau)\exp(-j\omega_k \tau)d\tau = \frac{1}{T_u^{\,2}}\int_0^{T_u}\int_0^{T_u} u(t)\overline{u(t-\tau)}\exp(-j\omega_k \tau)d\tau dt\,. \quad (18)$$

Introduce the substitution of variables $\tau = t - \eta$ and take into account the periodicity on η:

$$K_u(\omega_k) = \frac{1}{T_u^2}\int_0^{T_u}\int_0^{T_u} u(t)\overline{u(\eta)}exp(-j\omega_k(t-\eta))d\eta dt\ . \tag{19}$$

Where does follow:

$$K_u(\omega_k) = U(\omega_k)\overline{U(\omega_k)},\ \text{and}\ \left|U(\omega_k)\right| = \sqrt{K_u(\omega_k)}\ , \tag{20}$$

therefore

$$K_u(\omega_k) = \frac{U_0^{\ 2}Q(\omega_k)\overline{Q(\omega_k)}}{M^2}\sum_{n=0}^{M-1}\sum_{m=0}^{M-1}S(n)\overline{S(m)}exp(-j\omega_k \Delta T(n-m))\ . \tag{21}$$

Let $m = n - p$, since the sequence $S(n-p)$ and function $exp(-j\omega_k \Delta T)$ are periodic with index p with the period M, the summation limits can be rearranged:

$$K_u(\omega_k) = \frac{U_0^{\ 2}Q(\omega_k)\overline{Q(\omega_k)}}{M^2}\sum_{p=0}^{M-1}\sum_{n=0}^{M-1}S(n)\overline{S(n-p)}exp(-j\omega_k \Delta Tp)\ . \tag{22}$$

Using the value of the discrete autocorrelation function of the signal $S(n)$, we have:

$$K_u(\omega_k) = \frac{U_0^{\ 2}Q(\omega_k)\overline{Q(\omega_k)}}{M}\left(1 - \frac{1}{M}\sum_{p=1}^{M-1}exp\left(-j\omega_k \Delta T\ p\right)\right). \tag{23}$$

If $exp\left(-j\omega_k \Delta T\right) \neq 1$, then the internal sum can be calculated as sum of the geometric progression:

$$\sum_{p=1}^{M-1}exp\left(-j\omega_k \Delta T\ p\right) = \frac{exp\left(-j\omega_k \Delta T\right)\left(1 - exp\left(-j\omega_k \Delta T(M-1)\right)\right)}{1 - exp\left(-j\omega_k \Delta T\right)} = -1\ . \tag{24}$$

If $exp\left(-j\omega_k \Delta T\right) = 1$, then the internal sum equal to M. Thus:

$$\left|U(\omega_k)\right| = \frac{U_0\left|Q(\omega_k)\right|}{M} \times \begin{cases}\sqrt{M+1} & if\ k \neq zM \\ 1 & if\ k = zM\end{cases} = \frac{U_0\left|Q(\omega_k)\right|}{(2^N - 1)} \times \begin{cases}2^{N/2} & if\ k \neq z\,M; \\ 1 & if\ k = z\,M.\end{cases} \tag{25}$$

It can be concluded that with an increase in the size N of the shift register by one bit, the amplitude of the harmonics of the modulated signal decreases by about $\sqrt{2}$ times.

Determination of frequency characteristics of current through the lamp and voltage on the lamp

For simplicity a pseudo-linear model of a discharge lamp are used. In this model ohm resistance of the lamp $R_л$ "almost do not change" for the period T_u, and depend only on the RMS value of the current of the period T_u. Suppose that a circuit with a transfer function $W(\omega)$ is connected between the amplifier output and the lamp. In this case, the harmonics of current $I_л(\omega_k)$ and voltage $U_л(\omega_k)$ of the lamp are determined by the harmonics of the signal $u(t)$:

$$I_л(\omega_k) = \frac{W(\omega_k)U(\omega_k)}{R_л},\ U_л(\omega_k) = W(\omega_k)U(\omega_k)\ . \tag{26}$$

In this case, the amplitudes of the harmonics are defined as:

$$\left|I(\omega_k)\right| = \frac{U_0\left|W(\omega_k)\right|\left|Q(\omega_k)\right|}{R_л\ M} \times \begin{cases}\sqrt{M+1} & if\ k \neq zM, \\ 1 & if\ k = zM,\end{cases},\left|U_л(\omega_k)\right| = \frac{U_0\left|W(\omega_k)\right|\left|Q(\omega_k)\right|}{M} \times \begin{cases}\sqrt{M+1} & if\ k \neq zM \\ 1 & if\ k = zM\end{cases}. \tag{27}$$

Therefore, the signal on the lamp behaves similarly to the signal at the output of the amplifier and with increasing bit rate of the LFSR register by 2 bits, the amplitude of the voltage harmonics decreases by about 2 times.

Determination of power harmonics on a lamp

Since at least the instantaneous value of the power is limited, and the current and voltage after the transients are periodic functions with a period T_u then the power will be periodic, so it can also be formally decomposed into a Fourier series by frequencies. $\omega_k = \dfrac{2\pi\, k}{\Delta T\ M}$. Formally, we can write the Fourier series coefficients of lamp power $P(t) = u_л(t)\, i_л(t)$ at $[0, T_u]$.

$$P(\omega_k) = \frac{1}{T_u}\int_0^{T_u} u_л(t)\, i_л(t) exp(-j\omega_k t) dt\,. \tag{28}$$

Substituting the voltage and current representations in the form of a Fourier series and changing the order of integration/summation, we obtain:

$$P(\omega_k) = \sum_{n=-\infty}^{\infty}\sum_{m=-\infty}^{\infty} U_л(\omega_n)\, \overline{I_л(\omega_m)} \frac{1}{T_u}\int_0^{T_u} exp\big(jt(\omega_n - \omega_m - \omega_k)\big)\, dt\,, \tag{29}$$

if $\omega_n - \omega_m - \omega_k = 0$, the internal integral is equal to T_u, otherwise it is 0. So, we can write:

$$P(\omega_k) = \sum_{n=-\infty}^{\infty} U_л(\omega_n)\, \overline{I_л(\omega_n - \omega_k)}\,. \tag{30}$$

Substitute the representation of the voltage and current of the lamp, and changing the order of summation, obtain

$$P(\omega_k) = \frac{U_0^{\,2}}{R_л\, M^2}\sum_{n=-\infty}^{\infty} W(\omega_n)\, \overline{W(\omega_n - \omega_k)}\, Q(\omega_n)\, \overline{Q(\omega_n - \omega_k)} \sum_{m=0}^{M-1}\sum_{l=0}^{M-1} S(m)\, \overline{S(l)}\, exp\left(\frac{2\pi j}{M}\big(-nm + nl - kl\big)\right). \tag{31}$$

Let $l = m - p$, and $n = q + k$. Due to the periodicity of the function $\overline{S(m-p)} exp\left(\frac{2\pi j}{M}\big(k\, m - q\, p\big)\right)$ in the variable p with the period M, the summation can be performed in the range from 0 to $M-1$.

$$P(\omega_k) = \frac{U_0^{\,2}}{R_л\, M^2}\sum_{q=-\infty}^{\infty} W(\omega_{q+k})\overline{W(\omega_q)} Q(\omega_{q+k})\overline{Q(\omega_q)} \times \sum_{m=0}^{M-1}\sum_{p=0}^{M-1} S(m)\, \overline{S(m-p)}\, exp\left(\frac{2\pi j}{M}\big(k\, m - q\, p\big)\right). \tag{32}$$

An attempt to sum the internal sum showed that the values obtained for the same M depend not only on k and p, but also on the initial value in the shift register and on the type of characteristic polynomial, so the search for a "general formula" that would give harmonic values power in a simple closed form is doomed to failure.

The power harmonics estimation

Let us estimate the power components that correspond to the value of k, which is either equal to 0 or a multiple of M (i.e. $k = iM$). In this case: $exp\left(\frac{2\pi j}{M}\big(k\, m\, M\big)\right) = 1$, therefore, the sum (32) will be significantly simplified:

$$P(\omega_{iM}) = U_0^{\,2}\sum_{q=-\infty}^{\infty}\frac{W(\omega_{q+iM})\overline{W(\omega_q)} Q(\omega_{q+iM})\overline{Q(\omega_q)}}{R_л\, M^2} \times \sum_{p=0}^{M-1}\left(\sum_{m=0}^{M-1} S(m)\, \overline{S(m-p)}\right) exp\left(\frac{2\pi j}{M}\big(-q\, p\big)\right). \tag{33}$$

Given the value of the discrete autocorrelation function of the signal $S(m)$, we have:

$$P(\omega_{iM}) = U_0^{\,2}\sum_{q=-\infty}^{\infty}\frac{W(\omega_{q+iM})\overline{W(\omega_q)} Q(\omega_{q+iM})\overline{Q(\omega_q)}}{R_л\, M} \times \sum_{p=0}^{M-1} exp\left(\frac{2\pi j}{M}\big(-q\, p\big)\right) \times \begin{cases} 1 & if\ p = zM; \\ -\dfrac{1}{M} & if\ p \neq zM; \end{cases} \tag{34}$$

where *z* is an integer, that can be written as

$$P(\omega_{iM}) = U_0^{\,2}\sum_{q=-\infty}^{\infty}\frac{W(\omega_{q+iM})\overline{W(\omega_q)} Q(\omega_{q+iM})\overline{Q(\omega_q)}}{R_л\, M} \times \left(1 - \frac{1}{M}\sum_{p=1}^{M-1} exp\left(\frac{2\pi j}{M}\big(-q\, p\big)\right)\right). \tag{35}$$

We can find the interior sum as the sum of geometric progression and obtained

$$P(\omega_{iM}) = \frac{U_0^{\,2}}{R_л\, M^2}\sum_{q=-\infty}^{\infty} W(\omega_{q+iM})\overline{W(\omega_q)} Q(\omega_{q+iM})\overline{Q(\omega_q)} \times \begin{cases} 1 & if\ q = z\, M; \\ M+1 & if\ q \neq zM. \end{cases} \tag{36}$$

As can be seen from the formula, as the length of the sequence increases, the value obtained under the sign of the sum increases almost proportionally M, so the power will not go to 0 when increases the length of the shift register. However, unwanted power harmonics can be reduced by selecting a modulating sequence.

With the parameter $i = 0$, the zero harmonic of the power will be determined, which are corresponding to the average power on the lamp:

$$P(0) = \frac{U_0^{\,2}(M+1)}{R_л\, M^2} \sum_{q=-\infty}^{\infty} W(\omega_q)\overline{W(\omega_q)}Q(\omega_q)\overline{Q(\omega_q)} - \frac{U_0^{\,2}}{R_л M} \sum_{r=-\infty}^{\infty} W(\omega_{rM})\overline{W(\omega_{rM})}Q(\omega_{rM})\overline{Q(\omega_{rM})}. \tag{37}$$

Simulation result

The schematic diagram that corresponds to the above structural scheme and generates the signal described, was modeled in the MicroCAP environment (Fig. 5). The clock source X1 generates a signal with a period of 60 µs and a frequency of 16.6 kHz, which is used for general synchronization of the circuit.

The functional generator U1 generates a reset signal that enters the corresponding inputs of the shift registers. The counter D2 (74HC393) divides the frequency received from V1 by 8. The shift registers DD3, DD4, (74HC164) together with the logical elements DD5_1, form a linear feedback shift register. The logic elements DD1_3 and DD1_4 together with the multiplexer DD6 (74HC251) create a sequence modulation generator and form a modulated pseudorandom signal that enters the output amplifier U2. To the output of the amplifier is connected sequentially LC ballast and discharge lamp HL1 (which is modeled by a resistor HL).

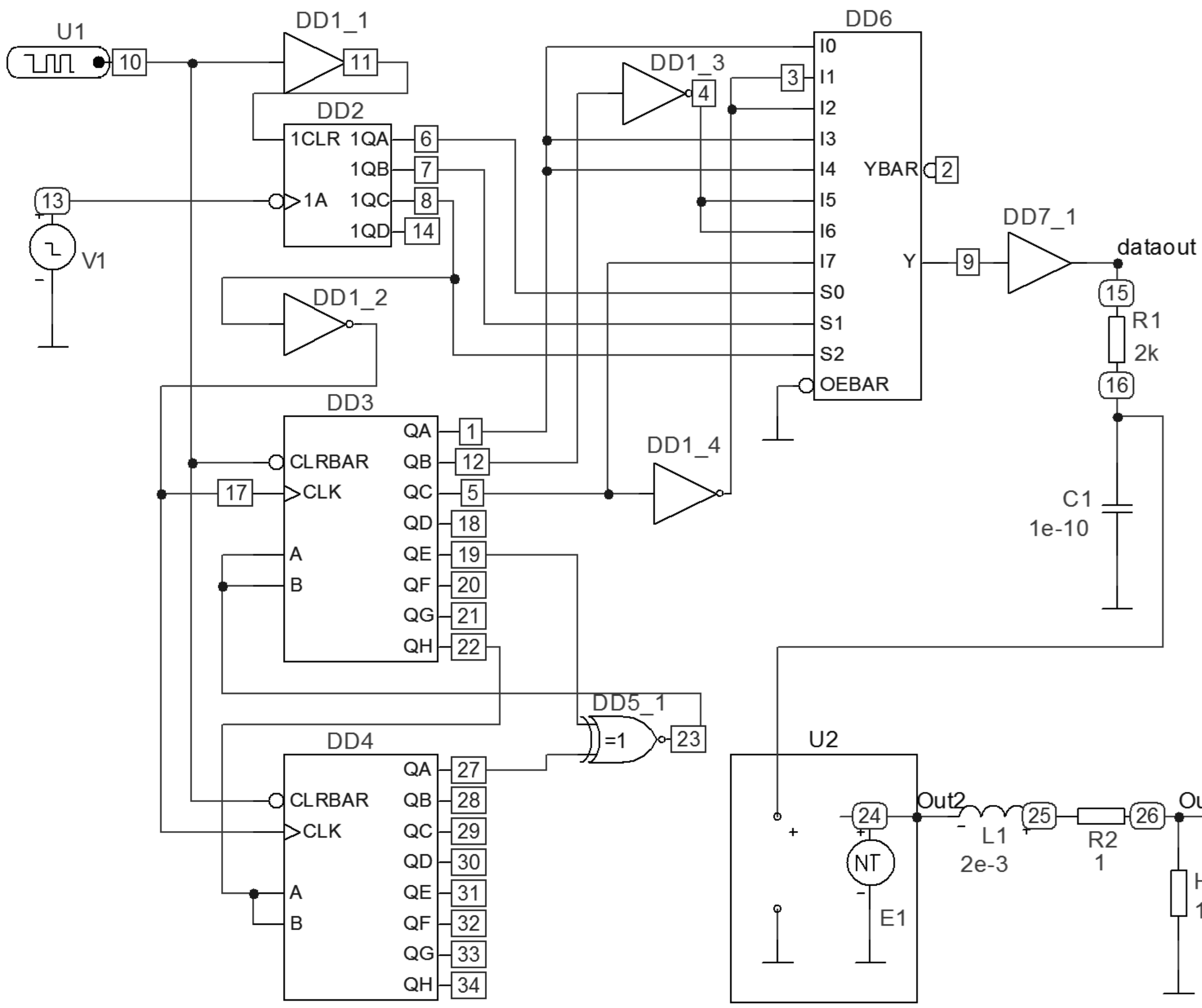

Fig. 5. Model of the offered scheme in MicroCAP environment

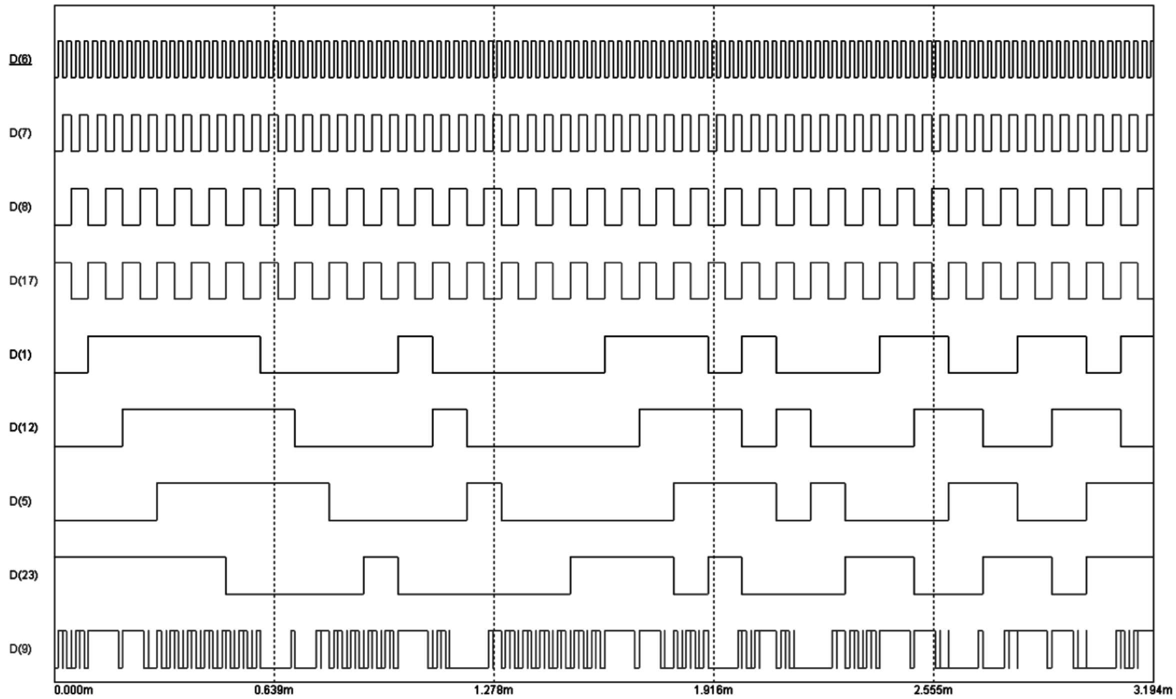

Fig. 6. Timing diagram of the main binary signals in the proposed circuit

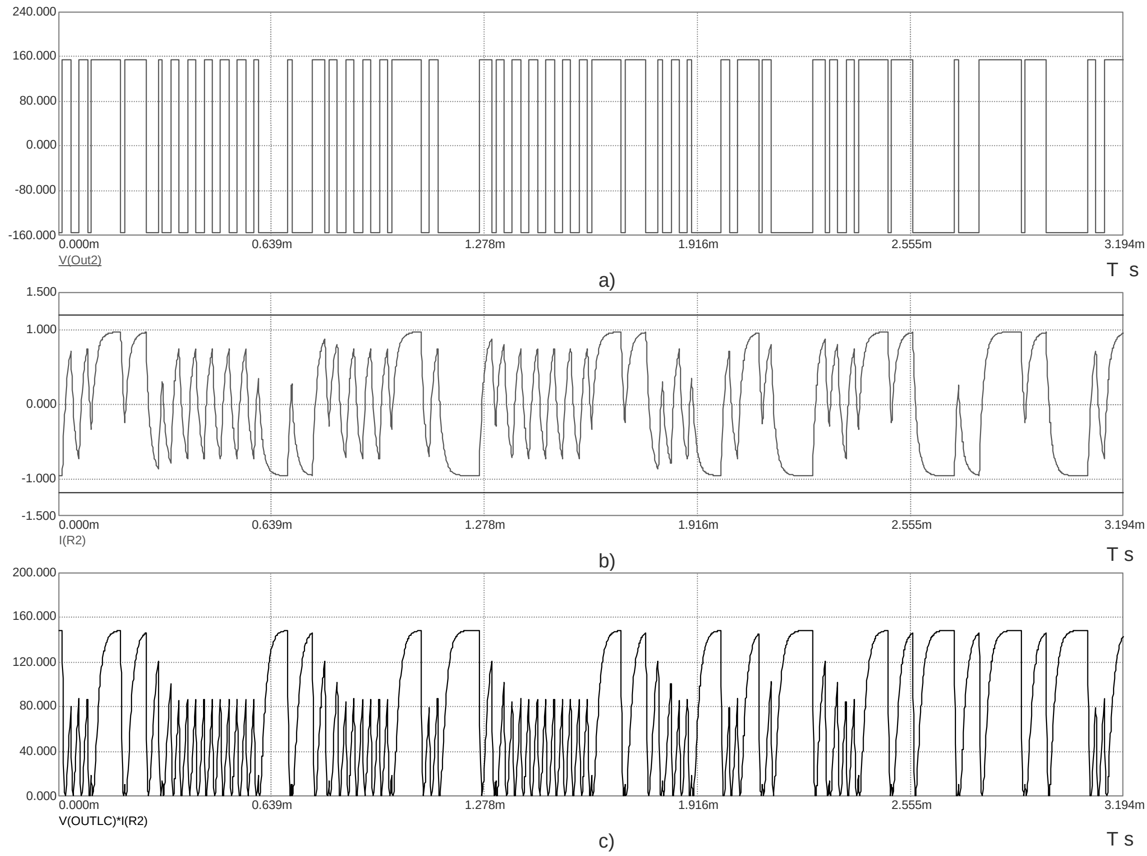

Fig 7. Fragment of the output signal: a) output voltage b) the lamp current c) instant power on the lamp

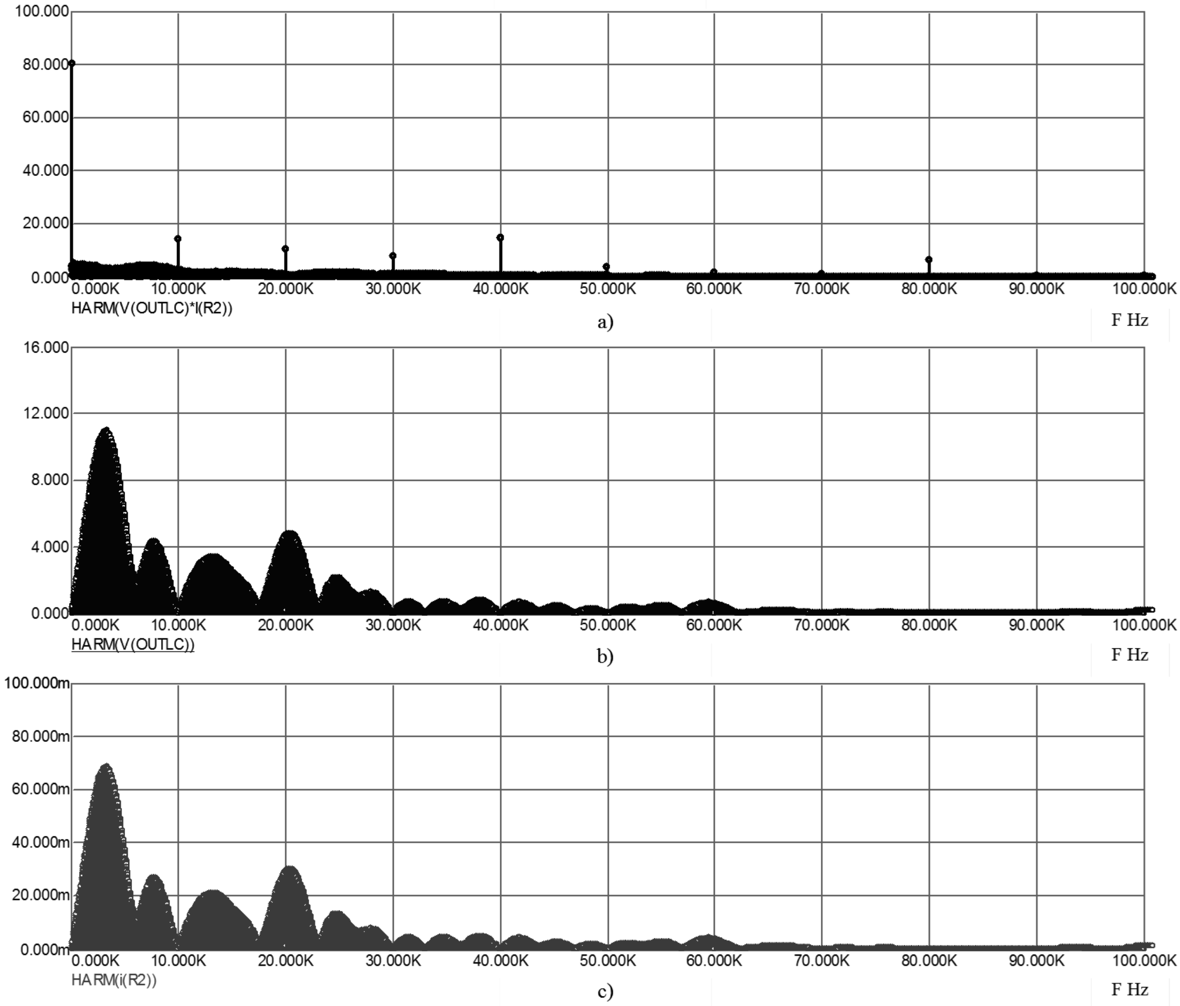

Fig. 8. Amplitude of harmonics of instant power of lamp (a), lamp voltage (b) and lamp current (c)

Results of simulation of the digital part of the scheme are shown in Fig. 6. Its shown the main digital signals of the proposed device: the counter DD2 output signals (D (6), D (7), D (8)), the synchronization signal of the LFSR register D (17), the LFSR register output signal D(1) and the output signal of the digital part of the circuit (D(9)). As follows from the figure, the output signal (D (9)), as it were, consists of low frequency and high frequency components alternating with each other in a pseudo-random order. The low frequency component gives the main switching of the current, while the high frequency component fills in the pauses of the low-frequency signal. The output signal of the analog part of the proposed device is shown in Figure 7.
The harmonics of the voltage across the lamp and the harmonics of the lamp power taken over the cycle of the shift register are shown in Figure 8. As can be seen from Figure 8, although the amplitude of the voltage across the lamp approaches hundreds of volts in the nominal operating mode, and the amplitude of the current to 0.8 A, the amplitude of each voltage harmonic does not exceed 12 V, and the amplitude of the current harmonic does not exceed 70 mA. In addition, while the average power - the power corresponding to the zero frequency is 80 W, each power harmonic does not exceed 17 VA.

Conclusions

The proposed principle of controlling high-pressure discharge lamps allows avoiding the phenomenon of acoustic resonance or reducing its effect. The main point of the scheme is to distribute energy over the largest possible frequency spectrum, which is being successfully implemented. As can be seen from Fig. 8, each harmonic of the output signal does not exceed the amplitude of 12 V in the frequency range from 0 to 100 kHz.
The circuit also reduces power harmonics at hazardous frequencies. Thus, the conditions for the occurrence of acoustic resonance are not created. Therefore, the use of a broadband signal based on the M-sequence automatically prevents acoustic resonance in high pressure discharge lamps.
The circuit implementation of the principle is quite simple and does not require the presence of a microcontroller. But, of course, the scheme for generating the necessary pseudo-random signal can be transformed into an algorithm and executed by some microcontroller.

Since the circuit is in no way tied to the properties of a particular type of lamp, there are no fundamental problems with the applicability of the proposed principle of power supply to discharge lamps of other types, for example, sodium.

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