FLIP-FLOP SENSOR CONTROLLED BY SLOW-RISE CONTROL PULSE

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Abstract

In this paper we deal with dynamic properties of the flip-flop sensor. Special attention will be paid to the condition of control by slow-rise segment of the control pulse and the derivation of the equivalent voltage. The results of the theoretical considerations are verified by simulations using SPICE, VERILOG, and a laboratory experiment.

Keywords

Flip-flop sensor, equivalent voltage, control pulse, non-electrical quantity, measurement.

1. Introduction

The key element of a flip-flop sensor is the switching circuit or the so-called elementary memory. Its simplest possible form is shown in Fig.1. It differs from the conventional elementary memory by its method of control. The control pulses are not applied to the base or gates of the switching circuit but the circuit is repeatedly connected to an ideal source of voltage or current. To be able to quantify the corresponding non-electrical signal, it is advantageous to use sensor elements in building the circuit. Instead of conventional load resistors it is possible to use e.g. piezoresistors, photoresistors, magnetoresistors etc. A similar situation arises in the case of transistors and inverters. It is better if they are phototransistors, magnetoresistors or transistors with multiple collectors which can be used in the circuit for the measurement of the magnetic field. For the quantification of the strain it is possible to use also transistors with a piezoresistive channel. By using ION SENSITIVE FET transistors it is possible to measure the pH of liquid media or to determine gas concentrations.

If needed however it is not necessary to stick to the principle that the elements in Fig. 1 should be sensometric. Sensometric elements can be connected to the circuit externally. They can be sensor bridges as well as active thermocouples, Hall probes etc.

But whichever of the above variants we choose we always have to stick to the principle that the measured non-electrical quantity will break the value symmetry of the inverters relative to the morphological symmetry axis passing through points **K** and **Z**. If for example identical phototransistors are used for the illumination measurement then the window of one of the bases is covered with aluminium foil which is done through a suitable technological process at the time of manufacture. Also in the strain measurement - if piezoresistors are used instead of load resistors they must be orthogonally oriented on the chip so that when the chip is deformed one of the resistors is strained longitudinally and the other transversally so that the value symmetry is broken during the strain.

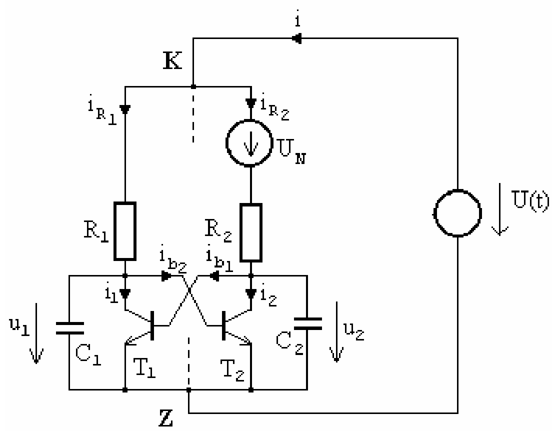

Fig. 1 Flip-flop sensor. Capacitances C_1 and C_2 represent parasitic capacitances of the transistors T_1, T_2.

Through the action of the measured non-electrical quantity the originally symmetrical transfer characteristics of the first inverter $u_1(u_2)$ and the second one $u_2(u_1)$ will be changed into asymmetrical ones. However, it can be compensated by a voltage $U_N = U_{NE}$ in such a way that by repeated connection to source $U(t)$ the 50% state is restored [1].

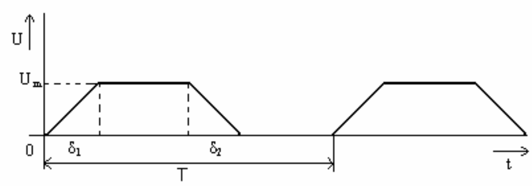

Fig. 2 Voltage control pulse.

It should be noted that in current and voltage control we also distinguish between the pulse with a steep or slow-rise segment of the control pulse (Fig.2). The control with a slow-rise segment of the control pulse is characterized by the ratio U_m / δ_1 being such that currents passing through the capacitors are negligible. The notion negligible is understood in relative sense. In practice condition is satisfied if δ_1, $\delta_2 >> R_1 C_1$ and δ_1, $\delta_2 >> R_2 C_2$ at the same time.

The flip-flop sensor can be described by the system of differential equations [4]

$$\frac{du_1}{dt} = \frac{R_2\left[U(t) - u_1 - R_1\phi_1\right]}{R_1 R_2 C_1} \equiv Q_1 \tag{1}$$

$$\frac{du_2}{dt} = \frac{R_1\left[U(t) - u_2 - U_N - R_2\phi_2\right]}{R_1 R_2 C_2} \equiv Q_2 \tag{2}$$

Quantities ϕ_1, ϕ_2 are defined as

$$\phi_1 = I_1 + I_2/\beta_2, \qquad \phi_2 = I_2 + I_1/\beta_1 \tag{3}$$

and

$$I_1 = i_{ES1}\exp(u_2/V_T), \quad I_2 = i_{ES2}\exp(u_1/V_T) \tag{4}$$

where β_1, β_2 are the current amplification coefficients, i_{ES1}, i_{ES2} are the saturation currents of bipolar transistors and V_T is thermal voltage

2. Triple point in the state plane

Specific to the control with a slow-rise control pulse is gradual appearance of singularities with gradual increase of the input voltage. The opposite situation occurs in case of control with a steep segment. When δ_1, $\delta_2 << R_1 C_1$ and δ_1, $\delta_2 << R_2 C_2$ the appearance of these singularities is not gradual but instantaneous. For $U<U_\alpha$ the singular point S_1 as an intersection point of the characteristics is stable. For $U=U_\alpha$ the intersection point of the transfer characteristics is the so-called triple point S_P [4]. If $U(t)>U_\alpha$, so the intersection points of the characteristics are points 0, S, 1. The points 0, 1 are stable, while the point S is unstable.

What is typical for the point S_P ?
Can be point S_p precisely analytically determined ?

Before answering these questions, it is necessary to clarify some questionable statements related to the circuit dynamics. It follows from the above conditions for control with a slow-rise segment that the currents passing through the capacitors are negligible compared to the transistor currents. This finding leads us to the notion non-dynamic circuit and makes one dubious about the use of expressions like stable or non-stable state. The notion negligible, however, is a relative one and in practice the capacitance currents are nonzero. Therefore it makes sense to consider the left-hand side of equations (1), (2). Also from the theoretical viewpoint the capacitor currents cannot be zero because that would correspond to a zero increment in the control pulse. This view of the problem makes it possible to put together the Jacobi matrix for given system (1), (2) relative to the treatment of the stability or instability of the given point. Consider a perfectly value-symmetric circuit. Then $R = R_1 = R_2$, $\beta = \beta_1 = \beta_2$, $i_{ES} = i_{ES1} = i_{ES2}$, $C = C_1 = C_2$. Let the point S_P be the triple point at the value of the input control pulse $U = U_\alpha$ and let $S_P = [U_S, U_S]$.

The Jacobi matrix in this case has the form:

$$\underline{J} = \frac{1}{RC}\begin{pmatrix} 1 + \dfrac{RI_\alpha}{\beta V_T} & \dfrac{RI_\alpha}{V_T} \\ \dfrac{RI_\alpha}{V_T} & 1 + \dfrac{RI_\alpha}{\beta V_T} \end{pmatrix} \tag{5}$$

where $I_\alpha = i_{ES}\exp(U_S / V_T)$. For the calculation of eigenvalues we have

$$\det\left(\underline{J} - \lambda \underline{E}\right) = 0 \tag{6}$$

where $\underline{E}$ is the unity matrix. It is obvious that for $U<U_\alpha$ the singular point S_1 as a intersection point of the characteristics is stable. For $U = U_\alpha$ the intersection point of the transfer characteristics is the so-called triple point S_P [4]. Triple point S_P is on landmark stability and instability, since eigenvalues $\lambda = 0$. For the resulting current $I = \phi_1 + \phi_2$ at point S_P we have:

$$I = 2I_\alpha\left(1 + 1/\beta\right) \tag{7}$$

and for the voltage

$$U_\alpha = RI_\alpha\left(1 + \frac{1}{\beta}\right) + V_T \ln\left(\frac{I_\alpha}{i_{ES}}\right) \tag{8}$$

where

$$I_\alpha = \frac{V_T}{R\left(1 - 1/\beta\right)} \tag{9}$$

Eqns. (7), (8), (9) were derived through (6), (3), (4) assuming $\lambda = 0$. Coordinate of S_P can be calculated analytically:

$$U_S = V_T \ln\left[\frac{V_T\left(1 + 1/\beta\right)}{Ri_{ES}}\right] \tag{10}$$

Relation (10) can be derived through (4) and (9).

In Fig. 3a T_1 represents the state trajectory and the time marks correspond to given time instants in Fig. 3b.

3. Formula for the equivalent voltage

In the case of value unsymmetry assume $S_P = [U_1, U_2]$ so that $U_1 \neq U_2$. The Jacobi matrix in this case has the form

$$\underline{J} = \begin{pmatrix} \dfrac{1}{R_1 C_1}\left(1 + \dfrac{R_1 I_2}{V_T \beta_2}\right) & \dfrac{I_1}{C_1 V_T} \\ \dfrac{I_2}{C_2 V_T} & \dfrac{1}{R_2 C_2}\left(1 + \dfrac{R_2 I_1}{V_T \beta_1}\right) \end{pmatrix} \tag{11}$$

where $I_1 = i_{ES1}\exp(U_2 / V_T)$, $I_2 = i_{ES2}\exp(U_1 / V_T)$. From the condition (6), for $\lambda = 0$, we have

$$1+\frac{R_2 I_1}{V_T\beta_1}+\frac{R_1 I_2}{V_T\beta_2}+\frac{R_1 R_2 I_1 I_2}{V_T^2}\left(\frac{1}{\beta_1\beta_2}-1\right)=0 \quad (12)$$

Despite that the elements of the Jacobi matrix depend on capacitors C_1, C_2 they do not affect the location of the point S_P in the state plane. In the equation (12) we included control with a slow-rise segment – the triple point S_P is meaningful in the functioning of the flip-flop sensor only if it is reached by a state trajectory, which corresponds to the control with a slow-rise segment of the control pulse.

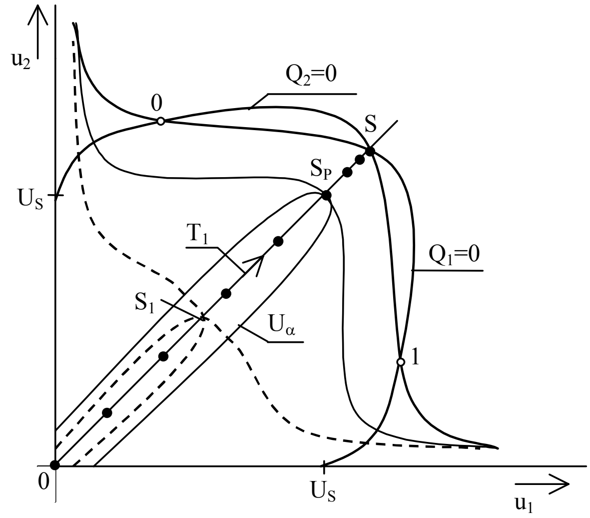

Fig. 3a Gradual appearance of the singular points

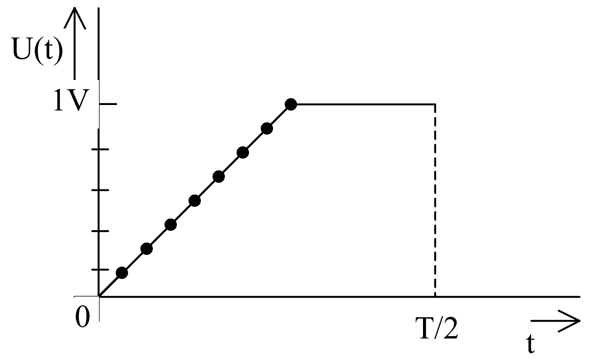

Fig. 3b Voltage control pulse in the half-period

Eqn. (12) is a necessary but not sufficient condition for the determination of S_P. Therefore a second equation has to be written, taking into account the equilibrium state at the value asymmetry. For $U_N = U_{NE}$, $U(t) = 0$ we must have $u_1(0) = 0$, $u_2 = -U_{NE}$. At equilibrium without the effect of noise we have $u_2 = u_1 - U_{NE}$. From 2nd Kirchhoff law

$$R_1\left(I_1+\frac{I_2}{\beta_2}+C_1\frac{du_1}{dt}\right)+u_1=U_{NE}+ \\ +R_2\left(I_2+\frac{I_1}{\beta_1}+C_2\frac{du_2}{dt}\right)+u_2 \quad (13)$$

Since the capacitor currents are negligible compared to the transistor ones and

$$u_1-u_2=U_{NE} \quad (14)$$

we obtain

$$R_1\left(I_1+I_2/\beta_2\right)=R_2\left(I_2+I_1/\beta_1\right) \quad (15)$$

This equation characterizes the equilibrium state in the case of value asymmetry. Solving the system (12), (15) we get

$$I_1=\frac{V_T}{R_1-R_2/\beta_1},\quad I_2=\frac{V_T}{R_2-R_1/\beta_2} \quad (16)$$

The equivalent voltage U_{NE} can be derived through (4), (14) and (16). The result is

$$U_{NE}=V_T\ln\left[\frac{\left(R_1-R_2/\beta_1\right)}{\left(R_2-R_1/\beta_2\right)}\frac{i_{ES1}}{i_{ES2}}\right] \quad (17)$$

The formula for U_α can by derived through (3),(4),(16).

$$U_\alpha=V_T\ln\left[\frac{V_T}{i_{ES2}\left(R_2-R_1/\beta_2\right)}\right]+\frac{V_T R_1\beta_1}{\beta_1 R_1-R_2} \quad (18)$$

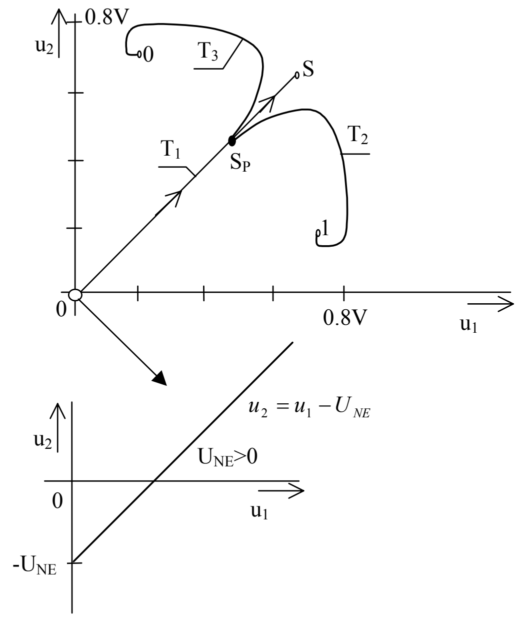

Fig. 4 Paths of singularities at equilibrium and the relation $u_2=f(u_1)$ in the neighborhood of the origin of the state plane.

What is the physical interpretation of the existence of the point S_P?

Taking the derivatives of the eqns. (4) and (13) it follows

$$R_1\left(dI_1+\frac{dI_2}{\beta_2}\right)+du_1=R_2\left(dI_2+\frac{dI_1}{\beta_1}\right)+du_2 \quad (19)$$

$$dI_1 = \frac{I_1}{V_T} du_2, \quad dI_2 = \frac{I_2}{V_T} du_1 \tag{20}$$

Eliminating dI_1 and dI_2 result in

$$\left[\frac{I_1}{V_T} \left(R_1 - \frac{R_2}{\beta_1} \right) - 1 \right] \frac{du_2}{du_1} = \frac{I_2}{V_T} \left(R_2 - \frac{R_1}{\beta_2} \right) - 1 \tag{21}$$

Obviously, (21) holds at the point S_P and at that point only for theoretically arbitrary ratio of the differentials du_2 / du_1. Hence, if affected by noise e.g., the voltages u_2, u_1 change in the neighborhood of S_P so that du_2 / du_1 < 1, then we observe a transition to state 1 (trajectory T_2). If du_2 / du_1 > 1, then the circuit assumes state 0 (trajectory T_3). Plausible cases are shown in Fig. 4 and it should be noted they are valid for U_{NE}>0. After theoretical analysis, we can conclude: point S_P is the point in which the decision is made about the transition to state 0 or 1 depending on du_2 / du_1, that is, the effect of noise, if $U_N = U_{NE}$. With the existence of white noise with zero mean value is connected the so-called 50 % state [1] – the number of transitions to state 0 is equal to the number of transitions to state 1, if $U_N = U_{NE}$.

4. Experimental results

The validity of the formula for the calculation of equivalent voltage was proved by simulation in SPICE, VERILOG and by a laboratory experiment. The parameter values of the bipolar transistors used are shown in Tab. 1 (SPICE). In Tab. 2 the equivalent voltages are obtained with the aid of the formula, simulation, and laboratory experiment. The value of the equivalent voltage can be determined in a laboratory experiment simply by gradually increasing of the symmetrisizing voltage until the so-called

IS = 10^{-16} A	RC = 10 Ω
BF = 100	CJC = 1 pF
ISC = 10^{-14} A	CJE = 1 pF
RB = 100 Ω	VJE = 0.75 V
RE = 1 Ω	VJC = 0.75 V

Tab. 1 Parameter values of transistors

R_2 = 6.778 [kΩ]	U_{NE} equation [mV]	U_{NE} SPICE [mV]	U_{NE} experiment [mV]
R_1 = 8.18	4.95	5.00	5.10
R_1 = 9.11	7.85	7.90	8.00
R_1 = 9.96	10.21	10.25	10.48
R_1 = 11.95	15.00	15.10	15.40

Tab. 2 Values of equivalent voltages

A 50 % state [1], when the value of the symmetrisizing voltage equals the equivalent voltage. A disadvantage of this method is the need of a manual change of the symmetrisizing voltage depending on the circuit's asymmetry. In practice it is better to use an auto-compensating system [3], [2] in which the equivalent voltage is set automatically depending on the asymmetry. In our case we used in the laboratory experiment an auto-compensating system with translation into time interval (the number of pulses N_1 read in). The system is shown in Fig. 5.

For a correct functioning it is necessary that $R_1 > R_2$ ($U_{NE} > 0$) holds. Depending on the asymmetry the voltage at the output of the integrator decreases into negative values according to formula $u_0 = -(U_{CC}\, t) / (R_3\, C_1)$.

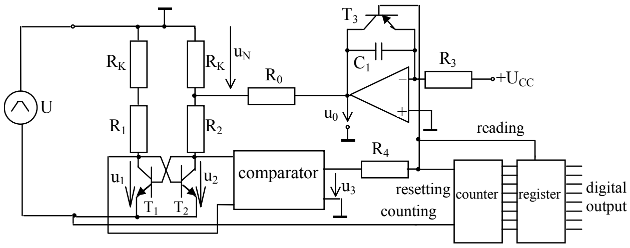

Fig. 5 Auto-compensating system

If $U_{NE} = -(U_{0E}\, R_K) / (R_0 + R_K)$, then the comparator starts working –negative voltage causes the transistor to switch – the voltage u_0 theoretically increases up to the zero value and since the symmetrisizing voltage is zero too, voltage u_0 decreases again into negative values. This repeats periodically so that to degree of asymmetry corresponds the number of pulses read in. This fact is expressed by Fig. 6.

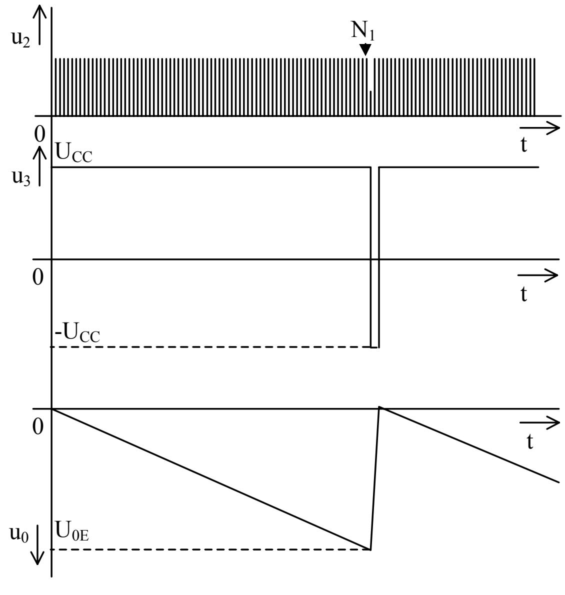

Fig. 6 Principle of the auto-compensating system functioning.

The flip-flop sensor was controlled by a voltage pulse according to Fig. 2, while δ_1, δ_2 = 18 μs, U_m = 1 V and T = 40 μs. From the obtained experimental results contained in Tab. 2 it follows that the inaccuracy of the equivalent voltage, calculated from relation (17) is less than 2.6%.

5. Conclusion

The aim of this paper was to show the properties of flip-flop sensor controlled by a slow-rise segment of the control pulse. From the formula (17) derived it is possible to study the effect of flip-flop asymmetry on the equivalent voltage. The validity of the formula for the calculation of equivalent voltage was proved by simulation in SPICE, VERILOG and by a laboratory experiment.

The asymmetry can be represented by the action of a given non-electrical quantity upon some of parameters of the flip-flop sensor. The method of measurement by the flip-flop sensor following from the existence of two stable states makes it possible to represent the equivalent voltage or a non-electrical quantity in a digital form without an extra AD converter.

From the formula (18) it is possible to study a dependence of the period of control pulse on the parameters of the flip-flop sensor, since for a correct detecting of state 'zero' or 'one' it is necessary that $U_m > U_\alpha$ holds, where U_m is amplitude of the control pulse (Fig. 2).

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30-TH ANNIVERSARY OF THE DEPARTMENT OF ELECTROMAGNETIC FIELD AT CTU PRAGUE

1. History

The Department of Electromagnetic Field at Czech Technical University in Prague (CTU) has been established thirty years ago. Although thirty years is less than 10% of CTU history, it is a point in time worth short review.

In 1971, departments of Theoretical Electrical Technology and Electromagnetic Wave Transmission were transferred into three new departments: Electromagnetic Field, Measurements and Circuit Theory.

The first head of the department emerging from this change was Prof. Haňka. Due to his age he retired in the school year 1974-75. Professor Tysl became the new head of the Department of the Electromagnetic Field, who resigned for the same reason as his predecessor in 1986. The next head of the department was professor Prokop. Unfortunately, his career ended as early as 1991 when he unexpectedly passed away. In 1991, doc. Novotný became the head of the department. He held the position until 1997, when he retired. Since 1997 the head of the department is doc. Miloš Mazánek.

Several hundred of students have completed their diploma theses at the department, and over fifty postgraduate students have reached the PhD. level there.

2. Current situation

The department is responsible for education of all students of the Faculty of Electrical Engineering in the field of Electromagnetic Field Theory. Besides that, the courses offered by the department include Microwaves, Antennas and Propagation, CAD, Optical Communications, Biological Effects of Electromagnetic Field, Medical Applications of Microwaves and more. Research is conducted in the same areas.

The department has paid a lot of interest to its laboratories. Microwave laboratory and an Anechoic Chamber for antenna and EMC measurement, opened 2000, are the latest in the row of laboratories.

Currently there are three full professors, five associate professors, two emeritus professors, twenty PhD students and other staff.

For more information and contacts see **http://www.cvut.cz/fee/k317**

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