PRACE Instytutu Maszyn Matematycznych PAN

Sprawozdania

Tom I Nr 3

THE CHARGE-CONTROL PARAMETERS OF ALLOY-DIFFUSED TRANSISTORS AND THEIR APPLICATION

by Tadeusz JANKOWSKI

P R A C E

Instytutu Maszyn Matematycznych
Polskiej Akademii Nauk

S P R A W O Z D A N I A

Tom I Nr 3

THE CHARGE-CONTROL PARAMETERS OF ALLOY-DIFFUSED TRANSISTORS AND THEIR APPLICATION

by Tadeusz JANKOWSKI

Warszawa 1964

Wydaje
INSTYTUT MASZYN MATEMATYCZNYCH PAN
Warszawa, ul. Koszykowa 79

Adres redakcji: Warszawa, ul. Koszykowa 79, tel. 28-37-29

Wykonano w IMM PAN. Nakład 150 egzemplarzy.
zam. 153/0/64.

Acknowledgements

The author wishes to thank Dr A.R.Boothroyd of Imperial College of Science and Technology, London, /where this work was performed in 1962 as the part of D.I.C. Dissertation/ for his supervision and helpful remarks, and appreciates the value of discussions on this paper with Mr D.Luke, M.Sc. and Mr P.Bylański, M.Sc.

The author acknowledges the support of the Polish Academy of Sciences, and particularly that of Prof. Dr L.Łukaszewicz of the Institute of Mathematical Machines, Warsaw, which enabled the performance of this work.

LIST OF PRINCIPAL SYMBOLS

b, B - base terminal or base region index

c, C - collector terminal index, capacitance

C_{Tc} - collector depletion layer capacitance

C_{Te} - emitter depletion layer capacitance

D_P - diffusion constant of holes

e, E - emitter terminal index

i - instantaneous value of current

I - D.C. or steady state value of current, inverse configuration

k - factor

$l \stackrel{def}{=} \frac{\tau_D}{\tau_o}$

m - factor

n - factor

N - fan-out, normal configuration

Q_B - charge associated with collector current

Q_{Bs} - excess charge in saturation

Q_E - charge associated with charging C_{Te}

$Q_R \stackrel{def}{=} Q_{Bs} + Q_B$ - total recombining charge

Q_V - charge associated with charging C_{Tc}

r, R - resistance

$r_{bb'}$ - extrinsic base resistance

r_{ecs} - emitter-to-collector resistance for the saturated transistor

t - time

V - voltage

W - base width

α_o - ratio of collector current to emitter current for active region

β_o - ratio of collector current to base current for active region.

β_s - "on demand" current gain

$\tau_B \overset{def}{=} \frac{Q_R}{I_B}$

$\tau_{Ba} = \beta_o \tau_c$

$\tau_{Bs} = \frac{Q_{Bs} + Q_B}{I_B}$

$\tau_c = \frac{Q_B}{I_c}$

$\tau_C = \tau_c + \tau_E$

$\tau_D \overset{def}{=} \frac{W^2}{2Dp}$

τ_E - emitter time constant /compare Appendix I/

τ_f - surface recombination time constant

τ_p - volume recombination time constant

τ_s - excess charge time constant

ω_α - alpha angular cut-off frequency

ω_T - gain-bandwith product

Instytut Maszyn Matematycznych
Sprawozdania
Nr 3
1964.06

621.382.3

THE CHARGE-CONTROL PARAMETERS OF ALLOY-DIFFUSED TRANSISTORS AND THEIR APPLICATION

by Tadeusz JANKOWSKI

Received December 1962

The alloy-diffused transistor feasibility of performing high speed operation is investigated in the paper. The charge-control parameters are considered the most suitable.

C O N T E N T S

INTRODUCTION

In the tests alloy-diffused transistors such as the OC170 and the ASZ20 were used. The charge control parameters are considered the most suitable. Also, it is shown that saturated operation is not to be recommended for fast computer circuits /using these transistors/. It is also worth emphasizing that the charge stored in diodes /whose number in relevant circuits usually exceeds several times the number of associated transistors/ may greatly influence the pulse properties of the designed circuits. This influence shows an additional advantage of charge control theory, being an approach that makes use of parameters which are uniform both for transistors and diodes.

The author considers the three-dimensional charge distribution as a prevailing factor determining the transistor behaviour in active region /wiggle effect/ and in saturation /the effective recombination time constant τ_{Bs}/. However, in saturation, the influence of the collector current level is still to be solved.

It is assumed that the reader is familiar with basic charge control theory /e.g. [2] or [6] / as well as with basic ideas of Boolean Algebra /e.g. [11] /. Basic computer circuits and more complex circuits would demand a more detailed description.

1. AN ATTEMPT AT EVALUATION OF CHARGE CONTROL PARAMETERS.

1.1. A c t i v e   r e g i o n - τ_c.

The concept of charge control parameters, developed by Sparkes and Beaufoy [1], [2], [3], [6] and others, has proved its usefulness. However, in spite of the rather logical and elegant picture presented, there are still a few problems to be considered from the physical point of view as well as from that of applications. Even if one assumes that the basic formula of charge control theory*/

$$Q_B = \tau_c \cdot i_c \qquad /1/$$

/where: i_c - instantaneous value of the collector current/ is immediately valid while switching**/ there still exist problems related to the so-called "wiggle effect" [6] as well as the relationship between charge control and "classical" small signal parameters /e.g. between time constants and cut-off frequencies/.

The wiggle effect suggests the possible lack of equality /2/

$$\frac{Q_B}{I_C} \neq \frac{d\,Q_B}{d\,i_c} \qquad /2/$$

i.e. that $\tau_c(t)$ varies during a transient. A detailed study of the wiggle effect is not the task of this paper. Therefore, having considered the wiggle effect as the result of charge propagation inside the transistor***/ let us formally rewrite /1/ as:

$$Q_B = \frac{\tau_D}{l}\, i_c \qquad /3/$$

where: $l = \frac{\tau_D}{\tau_c} = f(t)$

$$\tau_D = \frac{W^2}{2\,D_p} = \text{const}$$

---

*/Additionally Q_E and Q_V must be taken into account. Unless otherwise stated, the symbols are the same as in [6].

**/For alloy transistors it is to be true for $t \geqslant 0.24\ \frac{W^2}{2D_p} \cong 0.24\ \frac{1.22}{\omega_\alpha}$ [4], [6], [10].

***/Some authors even say about "the charge in the bushes" reflecting three - dimensional propagation phenomena.

The formula /3/ calls to mind the results of simplified one-dimensional theory [5], [9]

$$Q_1 = \tau_D \, i_c \qquad /4/$$

Let us consider the results of measurements and theory [6] showing disagreement between cut-off frequencies and the time constant τ_c:

$$\frac{1}{\omega_\alpha} < \frac{1}{\omega_T} < \tau_c$$

It is common opinion that fast transients /at least for common base configuration/ are relatively well described by ω_α. Besides, formulas /3/ and /4/ can suggest the possibility that the effective charge influencing the collector current is smaller than the total Q_B. Eventually the question arises how can τ_c be used for predicting the transient behaviour of a transistor if it differs from $\frac{1}{\omega_\alpha}$ and varies during the transient.

The time constant τ_c for fast transients is expected to be smaller than that measured by the Sparkes' method [2]. It is quite possible than in order to switch the transistor on, the knowledge of ω_α is needed. However, the charge stored and being removed while switching off, should be described by another "slow" τ_c.

Considering the recombination, let us study the semi-empirical formula /5/ where, for simplification, Q_E and Q_V are neglected in active region:

$$\frac{d\,Q_B}{dt} + \frac{Q_B}{\tau_B} = i_b \qquad /5/$$

where: $\tau_B = \beta_0 \tau_c$ – effective recombination time constant for active region [2], [3].

Assuming τ_B = const one obtains the general solution

$$Q_B(t) = \left[Q_o + \int_0^t i_b(t)\, e^{\frac{t}{\tau_B}} dt \right] e^{\frac{-t}{\tau_B}} \qquad /6/$$

where: $Q_o = Q_B(0)$.

The above formula involves the well-known exponential relationships for transients [9] where, for instance, ω_β takes the place of $\frac{1}{\tau_B}$. Unfortunately, it is highly probable that if the transistor exhibits the wiggle effect, the assumption that τ_B = const is not true.

For $t \ll \tau_B$ /neglecting recombination/ $e^{\frac{t}{\tau_B}}$ is approximately equal to 1 and therefore

$$Q_B(t) \cong Q_o + \int_0^t i_b(t)\, dt \qquad /7/$$

1.2. Case of nonlinear operation.

Let us define nonlinear operation as switching from cut-off to saturation and/or vice versa. This kind of operation is typical of two-value logic or bistable circuits. There are two major phenomena which must be taken into account in addition to the small-signal active-region operation, namely the influence of depletion-layer capacitancies /especially important for alloy-diffused transistors with small τ_c of 1-2 nsec/ and time-constant dependence on charge distribution. Besides, the charge distribution depends on both, region of operation and injection level; though solved for active region [7], [15] is not yet sufficiently known for saturation.

A. Switching from off to the active region.

The charge Q_E associated with charging the emitter depletion layer capacitance is the prevalent factor while performing the discussed operation, as ΔI_C /involving Q_B/ and ΔV_C /involving Q_V/ are considered small; Q_E can be treated as the charge needed to change voltage across the emitter-base junction. This charge is equivalent to modified time constant $\tau_C = \tau_c + \tau_E$ /compare Appendix I/ but, while going from the off state, τ_E varies and the picture is more complicated.

Let us assume the following relationship /see fig. 1/

$$Q_E = C_{Te} \cdot \Delta V_{eb'} \qquad /8/$$

While switching, the input voltage V_{eb} changes from the reverse value somewhat lower than 0.5 V /because of the limiting values for the OC170 and ASZ20/ up to the forward voltage somewhere between 0.2 V and 0.4 V. It is difficult to estimate*/ C_{Te} because of the change of sign of voltage across the emitter-base diode and eventually direct measurements of Q_E are more convenient than its predicting from /8/.

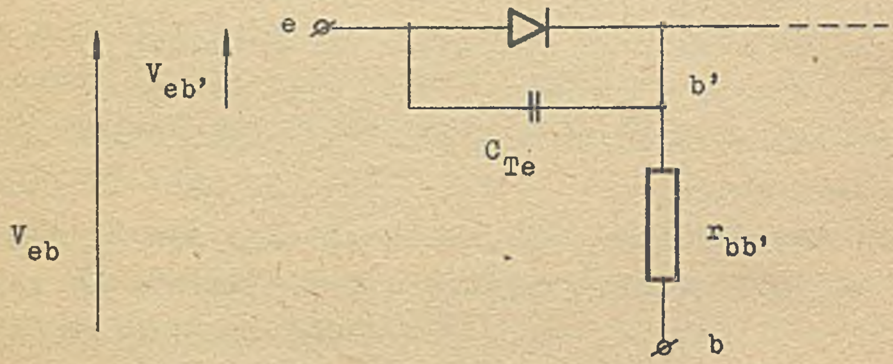


Fig. 1. Equivalent circuit for switching from the off state. The propagation phenomena are not yet taken into account.

*/ even the effective value

The proposed equivalent circuit /fig. 1/ is similar to that from [1] and has been confirmed by the measurements below /section 2.3/.

B. Crossing the active region.

In this case $Q_V = C_{To} \cdot \Delta V_{cb'}$ is needed besides Q_B which is described by /1/. Q_E is then much less important as $\Delta V_{eb'}$ is very small. One can predict Q_V following [6] by:

$$C_{To} = C_{tc1} \frac{2\left[-\phi - V_{cb1} - \sqrt{(\phi + V_{cb1})(\phi + V_{cb2})}\right]}{V_{cb1} - V_{cb2}} \qquad /9/$$

where

$$\Delta V_{cb'} \cong V_{cb1} - V_{cb2}$$

but, at least for the sake of uniformity of parameters, the direct measurements of Q_V would be recommended.

C. Charge in saturation.

Let us define saturation as the state in which the total recombining charge Q_R is greater than Q_B i.e.:

$$Q_R \stackrel{def}{=} Q_{Bs} + Q_B > Q_B \qquad /10/$$

It can be easily shown that the introduction of Sparkes' and Beaufoy's time constant τ_s for the excess charge Q_{Bs} from /10/ is synonymous with the following equation

$$\frac{Q_{Bs}}{\tau_s} + \frac{Q_B}{\beta_0 \tau_c} = I_B \qquad /11/$$

Equation /11/, having assumed a superposition of "normal" and "inverse" excess charge, does not describe physically the influence of charge recombination and distribution. Therefore, as the next approach, let us consider the steady state relationship extrapolated from equation /5/

$$\frac{Q_R}{\tau_{Bs}} = I_B \qquad /12/$$

Equation /12/ giving an "extrinsic" view of the transistor, describes the effective recombination i.e. the relationship between the total base current and charge. The time constant τ_{Bs} can be expressed by means of the time constant of volume recombination τ_p /this is for base region; however, there is possibly a component due to the recombination in the collector region/, the time constant of surface recombination τ_f and the factor of charge distribution k

$$\tau_{Bs} = \frac{\tau_p}{k + (1-k)\frac{\tau_p}{\tau_f}} \qquad /13\ a/$$

where

$$k = \frac{Q_1}{Q_R} < 1$$

Q_1 is an equivalent to that from /4/

$$Q_2 = Q_R - Q_1$$

$$\frac{Q_1}{\tau_p} + \frac{Q_2}{\tau_f} = I_B \qquad /13\ b/$$

And finally, before proceeding, let us consider the results obtained by the simple comparison of /11/ and /12/

$$\tau_{Bs} = \tau_s + \frac{I_C}{\beta_0 I_B} (\beta_0 \tau_c - \tau_s) \qquad /14/$$

The question arises which time constant is most representative and suitable from the point of view of both measurements and applications. The best proof of the complexity of this question is the comparison of the time constants used by different authors for the computation of the same storage time t_s:

1. Moll [9] - $\dfrac{\omega_N + \omega_I}{\omega_N \omega_I (1-\alpha_N \alpha_I)}$

2. Le Can [5]- $\dfrac{1/\omega_N + 1/\omega_I}{1/\beta_N + 1/\beta_I}$

   where index "N" means normal configuration and index "I" means inverse configuration.

3. Beaufoy [3] /for the total time t_{off}/ $\beta_0 \tau_c$

4. from eq. /12/ assuming τ_{Bs} = const τ_{Bs}

5. from eq. /11/ assuming τ_s = const τ_s

The author would suggest τ_{Bs} but the correctness of this opinion probably depends on the particular type of the transistor used.

Having assumed τ_s = const one can expect that τ_{Bs} varies as a function of I_B and I_C. However, this assumption has not yet been justified and τ_s can exhibit dependence on collector current /compare [16]/ as well as τ_{Bs} defined by /12/. The factor k of /13a/ suggests that τ_{Bs} varies only as far as the charge distribution depends on the collector current level.

D. "On demand" current gain β_s.

"On demand" current gain β_s is to be considered when a transient collector current is required and the transistor is in saturation [2].

$$\beta_s \overset{def}{=} \frac{\text{Available transient collector current}}{\text{Steady base current}}$$

The charge present in the base must be sufficient to satisfy the demands of a larger collector current and since generally τ_{Bs} can be either smaller or greater than the active region value of $\tau_B = \beta_0 \tau_c$, usually $\beta_s \neq \beta_0$.

If $\tau_{Bs} < \beta_0 \tau_0$, having switched the transistor from saturation to active region, one obtains an increasing charge due to the increased recombination time constant and the given constant base current. Therefore β_s is to be defined for the instant of switching, since later on the transistor will be able to deliver a greater collector current.

But if $\tau_{Bs} > \beta_0 \tau_c$, the charge in active region is decreasing due to the decreased recombination time constant and β_s must be defined for the worst case i.e. for the longest time t_n needed. The general expression for β_s has been computed in Appendix II.

$$\beta_s = \beta_0 + \left[\frac{\tau_{Bs}}{\tau_0} - \beta_0 \right] e^{\frac{-t_n}{\beta_0 \tau_c}} \qquad /15/$$

For $t_n = 0$ and $\tau_{Bs} = \tau_s$ one obtains the known relationship $\frac{\tau_s}{\tau_0}$ [2], [3]. If $\tau_{Bs} = \beta_0 \tau_0$, β_s is equal to β_0.

2. MEASUREMENTS ON THE OC170*/

2.1. General remarks.

The principle of measurement has been based on the Sparkes' method [2] but, mainly because of the lack of good variable capacitors and resistors, fixed components were used and input pulse voltages varied. Therefore, the most useful equations are /fig. 2/.

$$Q_{IN} = C_{B1} \cdot V_{1C} \qquad /16a/$$

and

$$\Delta I_C = \frac{\Delta V_C}{R_L} \qquad /16b/$$

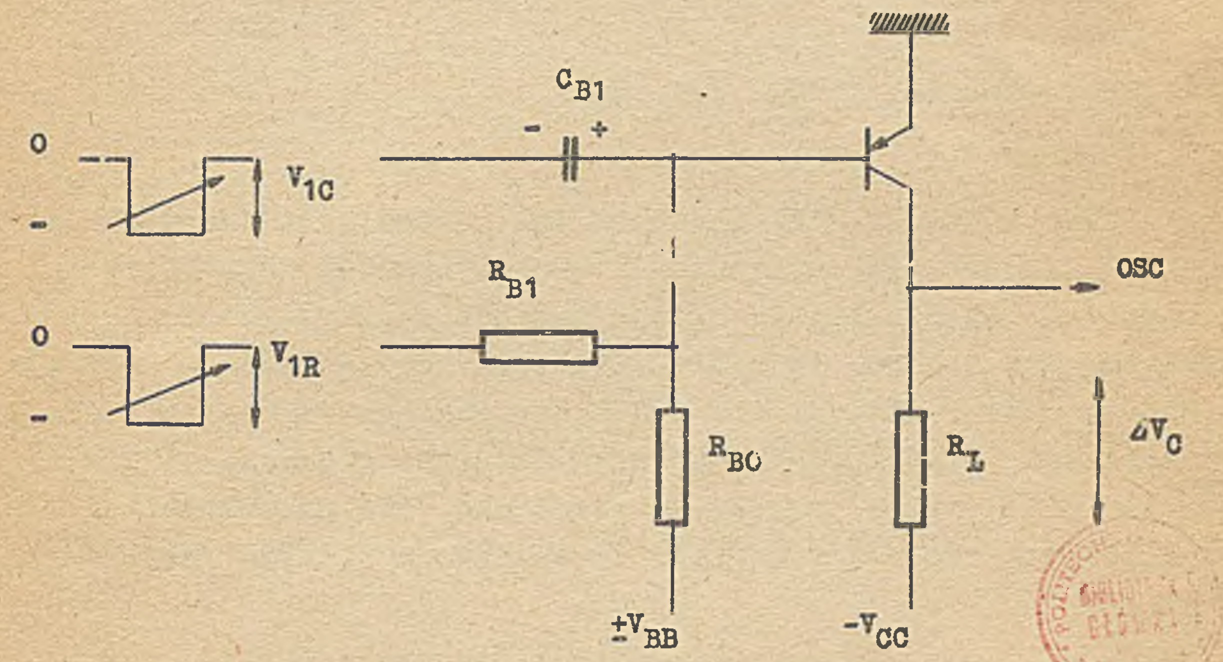


Fig. 2. The principle of measurement.

---

*/Some similar data on the ASZ20 are contained in [12]. The results obtained on П403А and П416А are of the same order.

Certain precautions were needed because of the limited maximum reverse emitter-base voltage /0.5 V for OC170/ and emitter reverse current /1mA/ as well as because of the apparently limited forward base current. Therefore, a 1kΩ resistor was sometimes introduced in series with C_{B1}; this was needed when using V_{1C} of the order of 10V.

A Tektronix 545 /N° 11337/ oscilloscope with a Type L plug-in unit /N° 100534/ was used. The total rise time of this instrument is of 12 nsec, the accuracy of the time base being not worse than 3% and the accuracy of the voltage measurements not worse than 5%.

The input voltage was obtained from an EMI Pulse Generator type 1 /Serial N° 101/ with a rise time of less than 10 nsec. A pulse width of 0.9 µs was obtained using an external delay line. Both positive and negative pulses with respect to ground were available. Unfortunately the top of the pulse was found to drop by 10%.

2.2. Measurements of Q_B and Q_V.

The practical circuit shown in fig. 3 had no adjustment of the D.C. pulse base current, as it was estimated that only a comparison between input voltage and the peak value of output waveform was needed. In other words, it was considered that the proportionality between the delivered charge and the collector current was "instantly" valid within the accuracy of Tektronix 545 /i.e. after approx. 12 nsec/ and that recombination could be neglected for periods less than ca 50 nsec /compare fig. 4/.

Table 1 shows some representative results for different conditions of measurements. First of all the collector current increment $\Delta I_C = \frac{\Delta V_C}{R_L}$ was varied and later on the initial collector current, I_C /including the case of a reverse biased emitter-base junction/.

For all the measurements the collector voltage increment ΔV_C was kept constant and equal to 5V.

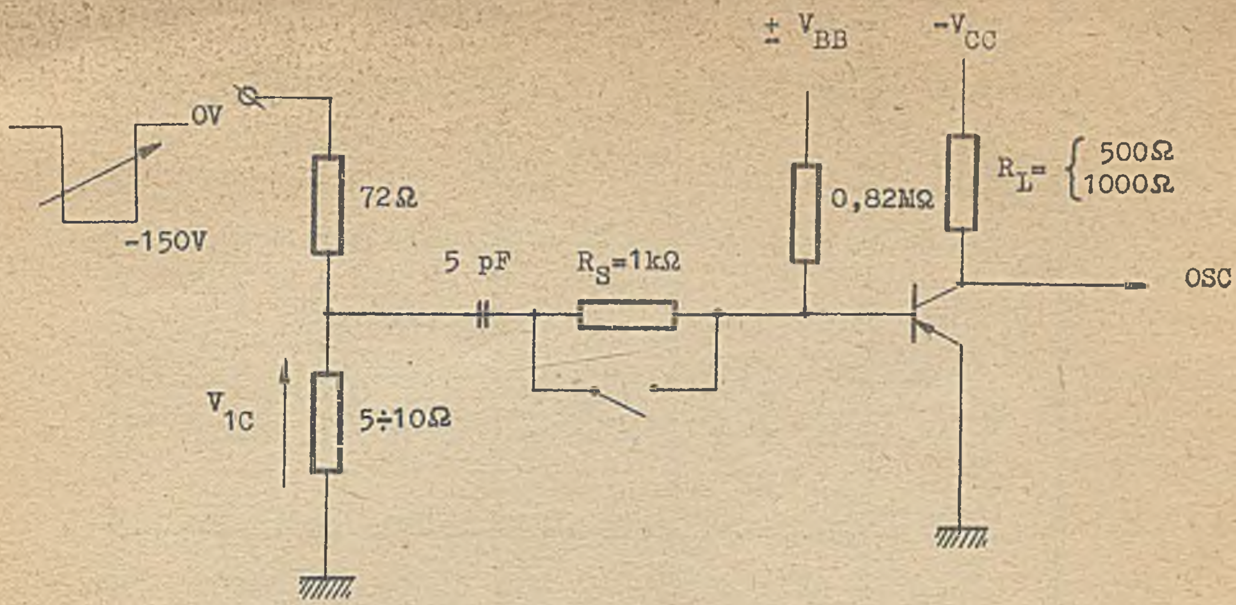


Fig. 3. The practical circuit for measurements of Q_E, Q_B, and Q_V.

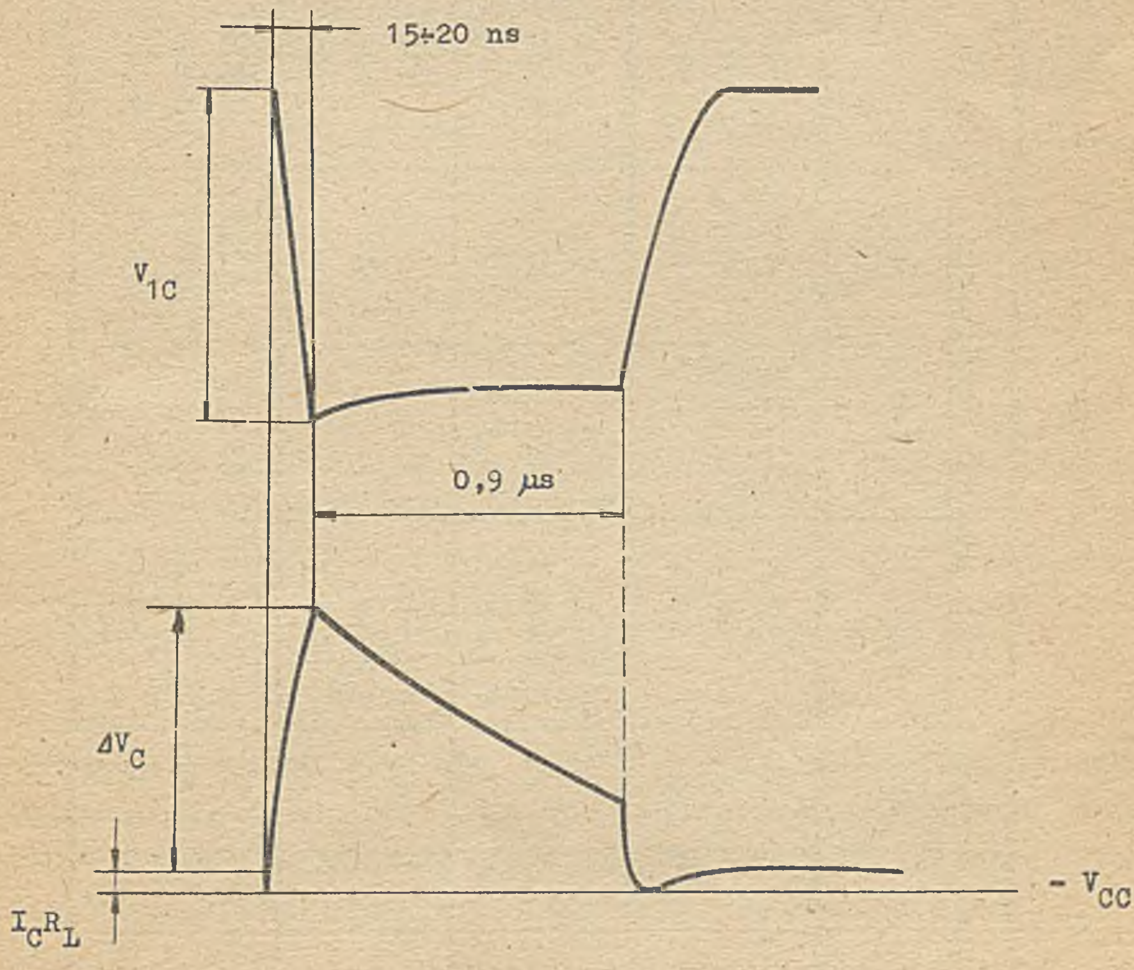


Fig. 4. Input and output waveforms.

Table 1

The measurements of Q_E, Q_B and Q_V

Case	Transistor No	The initial conditions V_B [V]	I_C [mA]	$-V_{CC}$ [V]	R_L [Ω]	ΔI_C [mA]	$V_{1C}-\Delta V_{b}$ [V]	Q_{IN} [pC]
A.	1.	+0.4	I_{CO} approx.	-6.5	500	10	9.8	49
	2.	+0.4	I_{CO} approx.	-6.5	500	10	9.8	49
	3.	+0.4	I_{CO} approx.	-6.5	500	10	8.3	~ 42
B.	1.	approx. -0.3	0.5	-7	1000	5	5.4	27
	2.	approx. -0.3	0.5	-7	1000	5	5.4	27
	3.	approx. -0.3	0.5	-7	1000	5	4.5	~ 23
C.	1.	approx. -0.3	1	-7	500	10	6.8	34
	2.	approx. -0.3	1	-7	500	10	6.8	34
	3.	approx. -0.3	1	-7	500	10	5.8	29
D.	1.	approx. -0.3	6	-12.5	1000	5	5.2	26
	2.	approx. -0.3	6	-12.5	1000	5	5.2	26
	3.	approx. -0.3	6	-12.5	1000	5	4.3	~ 22

C o n c l u s i o n s

I. The comparison of B. and C. /table 1/ shows that the charge needed to increase the colleotor current in the active region is relatively small for a fixed ΔV_C. Indeed, approximately the average value of

$$\tau_c = \frac{Q_{IN(C)} - Q_{IN(B)}}{\Delta I_{C(C)} - \Delta I_{C(B)}} = 1.2 - 1.4 \text{ nsec}$$

which for $\Delta I_C = 10\text{mA}$ corresponds to $Q_B = 12 - 14$ pC.

II. The comparison of B. and D. /having in mind C./ suggests that because of a very slight dependence on the initial collector current, the major input charge is needed to charge the capacitor C_{Tc} /Q_V/ and that Q_E may be neglected in the active region.

$$Q_V = Q_{IN(C)} - \Delta I_{C(C)} \tau_c = C_{Tc} \cdot \Delta V_C = 17 - 20 \text{ pC}$$

III. The comparison of A. and C. indicates the presence of the important charge Q_E needed for switching the transistor from the off state. Q_E can be estimated as the result of subtraction of $Q_{IN(C)}$ from $Q_{IN(A)}$

$$Q_E \cong Q_{IN(A)} - Q_{IN(C)} = 13 - 15 \text{ pC}$$

2.3. The comparison between charge and current drive

In practical circuit the charge drive involves high, but finite base current $i_b(t)$ exponentially decreasing with a time constant which is much smaller than the recombination time constant τ_B. The charge delivered Q_{IN} equals

$$\int_0^t i_b(t)\, dt$$

and it is quite probable that for $t \ll \tau_B$ the waveform of $i_b(t)$ is not so important. In particular it can just as well be a step function, providing the proper quantity of charge is delivered.

This is synonymous with common emitter current drive and from eq. /7/ one can obtain the formal approach for $t \ll \tau_B$

$$\frac{Q_{IN}}{\tau_c} \cong \frac{I_B}{\tau_c} t = i_c(t)$$

Fig. 5 shows the circuit for the current drive measurements while table 2 contains comprehensive results both for current and charge /i.e. that of fig. 3/ drive for the same transistor.

For the current drive the input pulse width was limited to 110 nsec. The collector current increased linearly after the characteristic delay while switching from the cut off state. It is considered that this delay was due to the time needed to charge the capacitor $C_{Te}(Q_E)$ i.e. to the time needed to switch on the emitter-base junction /see fig. 1/. Almost the same delay was obtained for the relatively "slow" alloy transistors OC44 and V6/R8.

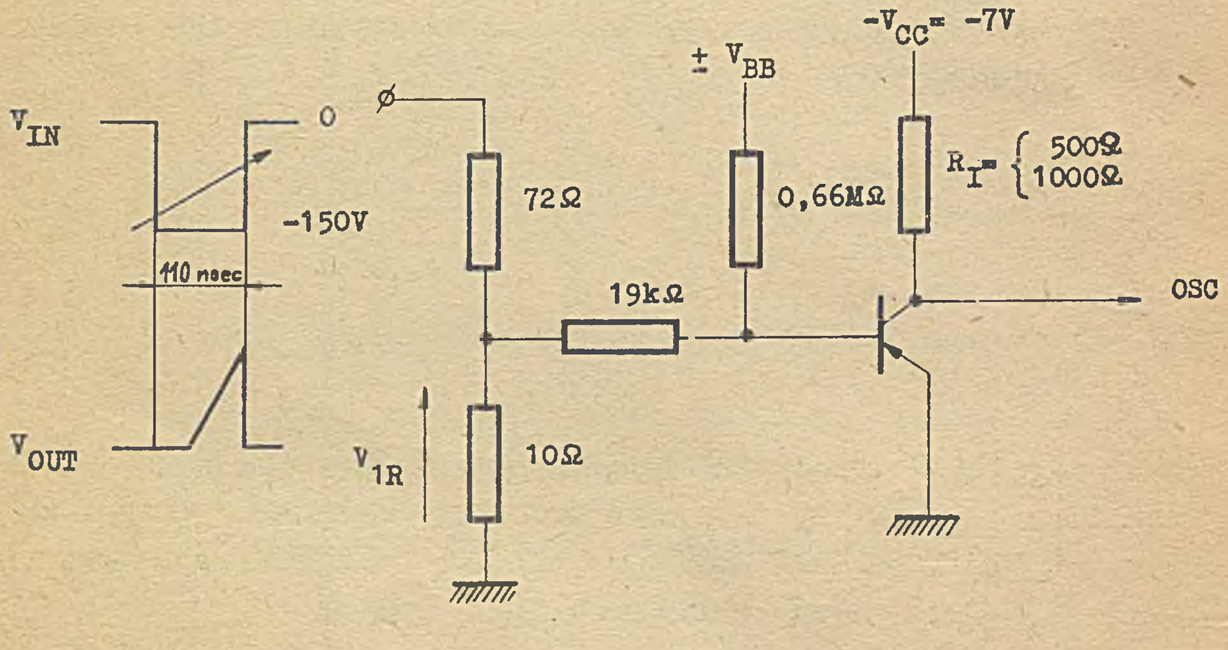


Fig. 5. Current drive measurements.

The comparison between charge and current drive

Table 2

Case No.*/	Conditions of switching		Current drive		Charge drive $R_S = 0$		Charge drive $R_S = 1k\,\Omega$	
	V_B / I_C	R_L [Ω]	V_{1R} [V]	Q_{IN} [pC]	$V_{1C}-\Delta V_B$ [V]	Q_{IN} [pC]	$V_{1C}-\Delta V_B$ [V]	Q_{IN} [pC]
1.	+0.2V / 1mA	500	5.2	30	**/ 2.3	11.5	4.5	22.5
2.	+0.2V / 1mA	1000	5	29	**/ 3.2	13.2	4.7	23.5
3.	+0.2V / 5mA	1000	7.6	44	8.5	42.5	8.5	42.5
4.	+0.2V / 10mA	500	8.8	51	9.5	47.5	9.5	47.5
5.	1 / 3mA	1000	1.5	8.7	1.8	9	1.8	9
6.	3 / 5mA	1000	1.5	8.7	1.8	9	1.8	9
7.	1 / 5mA	1000	3	17.4	3.8	19	3.6	18
8.	1 / 5mA	500	1.9	11	2.4	12	2.2	11
9.	5 / 10mA	500	2.3	13.3	3	15	2.8	14
10.	1 / 9mA	500	3.8	22	WAS NOT MEASURED		WAS NOT MEASURED	
11.	1 / 10mA	500	4.2	24.3	WAS NOT MEASURED		WAS NOT MEASURED	
12.	0 / 10mA	500	5.2	30	?	?	5.5	27.5
13.	+0.2V / 10mA	500	9	52	?	?	11.5	57.5
14.	1 / 5mA	500	1.8	10	?	?	1.9	9.5

*/ for the same transistor

**/ exhibits the wiggle effect

?/ difficulties of switching occured.

Conclusions

The measurements show good agreement /with small exceptions/ for different kinds of drive as well as for switching on and off /cases 4. and 13., 8. and 14./ and partial drive /cases 5., 6. and 7. or 8., 9 and 11/. Certain difficulties occurred only for "pure" charge drive.

The Q_E which is almost equal to Q_{IN} for cases 1 and 2, differs from that of table 1.

The comparison of cases 7 and 10 gives the approximate value of τ_c = 1.15 nsec and C_{Tc} = 2.6 - 3 pF. For purposes of computation a constant τ_c was assumed and Q_E neglected.

2.4. The charge in saturation

The charge in saturation was measured for different base and collector currents. Two cases were considered: a/ crossing the boundary between the saturation and active region /removing of Q_{Bs}/ and b/ the total charge Q_{OFF} needed to switch the transistor off. The latter also included Q_V and Q_E. Table 3 contains the results obtained for a single transistor but similar data were obtained for two others. The results for the case I_B = 60 µA, $I_C \cong 0$ were in good agreement with those obtained by the method [17] of measuring

$$\int_0^t i_c \, dt \quad \text{for} \quad i_e = 0.$$

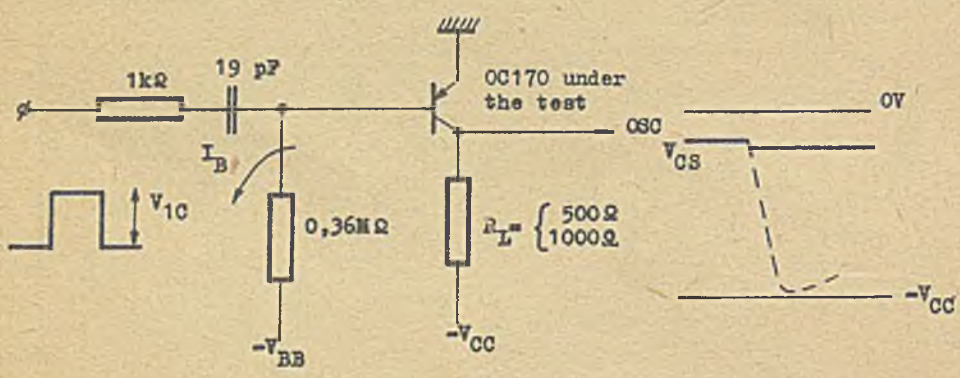


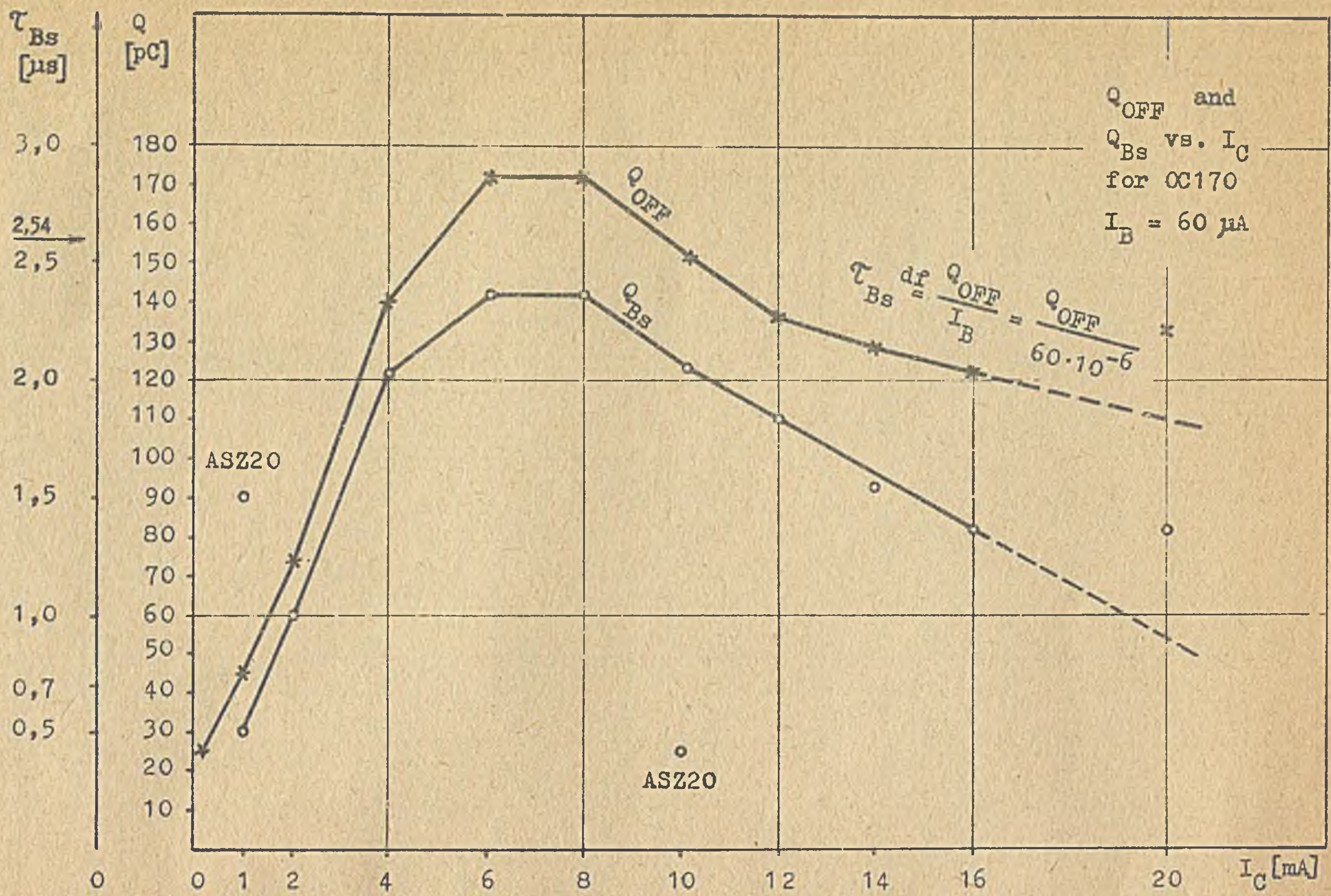
Fig. 6. Measurements of the charge in saturation.

Charge in saturation

Table 3

I_B [μA]	I_C [mA]	V_{CC} [V]	R_L [Ω]	V_{1Cs} [V]	$Q_{Bs} = V_{1Cs} \cdot 19$ pF [pC]	V_{1COFF} [V]	$Q_{OFF}=V_{1COFF} \cdot 19$ pF [pC]	$\tau_{Bs} = \frac{Q_{OFF}}{I_B}$ [μs]
100	10	approx.6	500	15	285	16	305	3.05
80	1	1.2	1000	2	38	3.1	59	0.74
	10	approx.6	500	8.4	160	11	209	2.6
	0.1		500	-	-	1.4	26.5	0.44
	1	1.2	500 /1000/	1,6	30	2.35	45	0.75
	2		500	3.2	61	3.8	72	1.2
	4		500	6.4	122	7.4	140	2.33
60	6		500	7.5	142	9	171	2.85
	8		500	7.5	142	9	171	2,85
	10	approx.6	500	6.5	124	8	152	2.54
	12		500	5.8	110	7.2	137	2.28
	14		500	4.9	93	6.8	129	2.15
	16		500	4.4	83	6.4	122	2.02
	20		500	4.4	83	7	133	2.22
40	1	1.2	1000	1	19	1.8	34	0.85
20	1	1.2	1000	0.5	9.5	1.2	23	1.15
10	1	1.2	1000	-	-	0.74	14	1.4

The plot of Q versus I_C for I_B = 60 μA is shown in fig. 7. Two relevant values for ASZ20 are enclosed



τ_{Bs} [µs]
Q [pC]
3,0
2,54
2,5
2,0
1,5
1,0
0,7
0,5
0
180
170
160
150
140
130
120
110
100
90
80
70
60
50
40
30
20
10
0
1
2
4
6
8
10
12
14
16
20
I_C [mA]
Q_{OFF}
Q_{Bs}
ASZ20
ASZ20
Q_{OFF} and
Q_{Bs} vs. I_C
for OC170
I_B = 60 µA
$\tau_{Bs} \overset{df}{=} \frac{Q_{OFF}}{I_B} = \frac{Q_{OFF}}{60 \cdot 10^{-6}}$

Fig. 7.

Conclusion

The most important point emerging from the above described is the dependence of the saturation time constant τ_{Bs} on the collector current. /Dependence on the base current is smaller and rather more understandable/. First of all one learns that τ_{Bs} is greater than the active region value $\beta_0\tau_c$ which corresponds to the catalogue data $\beta_{typ} = 100$ /usually $\beta_0 \leq 300$/ and $\tau_c = 1.2 - 1.4$ nsec. This is easily explained if one considers the transistor model with two time constants τ_p and τ_f -eq. /13/. The alloy-diffused type transistor, like the OC170, has an unsymmetrical geometry with its huge collector-base junction area and collector body. Therefore, in saturation the factor $k = \frac{Q_1}{Q_R}$ increases rapidly, and equation /14/ shows that it is the equivalent to $\tau_{Bs} > \tau_{Ba} = \beta_0\tau_c$. It means that the part played by volume recombination τ_p can be much greater in saturation than in the active region and, since $\tau_p \gg \tau_f$, the effective time constant τ_{Bs} should be greater than that corresponding to the active region.

The variation of τ_{Bs} with collector current is much more difficult to be explained and would probably demand a comprehensive consideration of the charge distribution when treating the transistor as a device with distributed parameters for the particular geometry involved. Besides, it is easy to find out that the variability of

$$\tau_s = \frac{Q_{Bs}}{I_B - \frac{I_C}{\beta_0}}$$

is of the same order.

Anyway, as Q_{OFF} /table 3/ is much greater than Q_{IN} /table 2/, operation in saturation cannot be recommended unless it is pulse operation /short periods in saturation/ or operation under very low collector currents.

2.5. "On demand" current gain β_s.

The main purpose of this section is the verification of equation /15/. The measurements were performed for the same transistor as in sec. 2.4. whence some of the data used for computation were derived. The circuit and relevant waveforms are shown in fig. 8. According to definition, "on demand" collector current was regarded as that for which the transistor emerges from saturation.

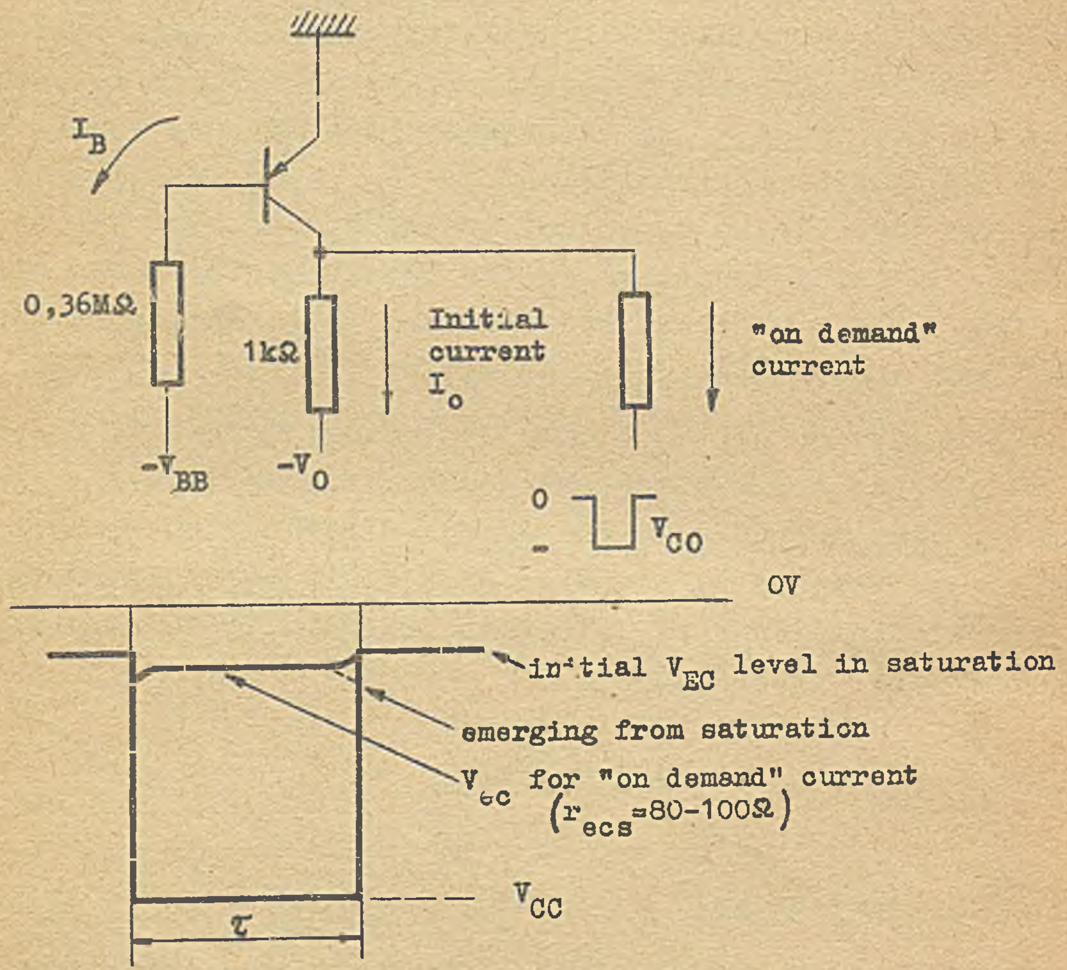


Fig. 8. "On - demand" measurements.

Table 4 contains the results; possible errors may be due to the variability of β_0 and β_s as functions of time and temperature.

"On demand" measurements

Table 4

Conditions	Measured $I_{on\ demand}$	$\beta_s = \frac{I_{on\ demand}}{I_B}$	β_s expected /eq.15/
I_B = 60 μA I_o = 1 mA τ = 0,9 μs	27mA	450	430
I_B = 60 μA I_o = 10mA τ = 0.9 μs	31.5 mA	525	610
I_B = 30 μA I_o = 1 mA τ = 0,9 μs	13.7 mA	456	430
I_B = 30 μA I_o = 1 mA τ = 0,1 μs	39 mA	1300	770

Finally let us compare results for I_B = 20 μA I_o = 1 mA and = 0,9 μs for some sampled OC170 transistors:

Sample	1	2	3	4	5	6
β_s	300	750	150	450	450	80

3. SOME PROBLEMS OF BASIC COMPUTER CIRCUITS

3.1. Introduction.

Although the application of transistors in contemporary computers is widespread, the designer still meets considerable difficulties due to numerous and often conflicting demands which notably include:

a/ good reliability over a wide range of ambient temperature;

b/ large tolerances of the components used;

c/ low cost of the circuitry as well as of power supplies and clocks;

d/ high speed;

e/ simple rules of interconnections between outputs and inputs;

f/ simplicity of packaging;

g/ ease of fault-finding and marginal checking and others, as for instance, the threshold for noise etc.

The problems mentioned above are sometimes hardly measurable and to a certain extent evaluated qualitatively, expecially when the most suitable technique and the most flexible set of the elementary "bricks" are considered.

First of all three points should be considered in detail:

1/ Fan - in

2/ Fan - out

3/ Delay per unit limiting the speed.

Moreover, the choice of transistor is influenced by its cost and delivery terms.

3.2. Transistor operation.

The main problem is to decrease Q_{IN}, since Q_{IN} equals $\int_0^t i_b \, dt$ and determines both the switching times /delay per stage/ and the input currents of the majority of transistorized circuits /i.e. indirectly it determines fan-out/. Let us consider the switching-on process when $Q_{IN} = Q_E + Q_B + Q_V$. To a first approximation the input current

$$I_{IN} = m \frac{Q_{IN}}{t_d}$$

/where: m = a factor greater than unity, t_d = switching time/, and the output current $I_{OUT} = n \frac{Q_B}{\tau_o}$ /where n = a factor less than unity/.

Therefore the fan-out factor can be written as:

$$N = \frac{I_{OUT}}{I_{IN}} = \frac{n\, t_d}{m\, \tau_o} \left(\frac{1}{\frac{Q_E}{Q_B} + 1 + \frac{Q_V}{Q_B}} \right) \qquad /17/$$

In formula /17/, t_d and τ_o are fixed for the assumed speed and the transistor used, the decreasing of Q_V is limited by the required output voltage swing, and the maximum Q_B is defined by the limitation of the collector current. Finally real progress can be achieved only by decreasing Q_E which plays almost the main role for the OC170 /compare sections 2.2. and 2.3/.

Fig. 9 shows the solution proposed by Miles [12] where the transistor is held either slightly conducting or heavily conducting and is never switched off. Since the emitter-base junction is still forward biased, Q_E is reduced to a small fraction of Q_B and, besides, there is no question of the reverse emitter-base break-down. The initial current I_K can be of the order of 0.1 - 0.3 mA compared with an emitter current of 6-10 mA. Using current-steering technique the defined current I_E flows in either the transistor T1 or an additional transistor /or diode/ according to the input voltage levels.

According to section 2.4. operation in saturation would be avoided by means of a clamping diode [11]. This implies a collector current well defined by the transistor. Otherwise, the transistor could be damaged, since out of saturation the external circuit does not limit the collector current. Therefore, the emitter current I_E would be established, as fig. 9 suggests, because for well defined I_B the spreading and variability of $\beta_0 = \frac{I_C}{I_B}$ can cause undesirable effects.

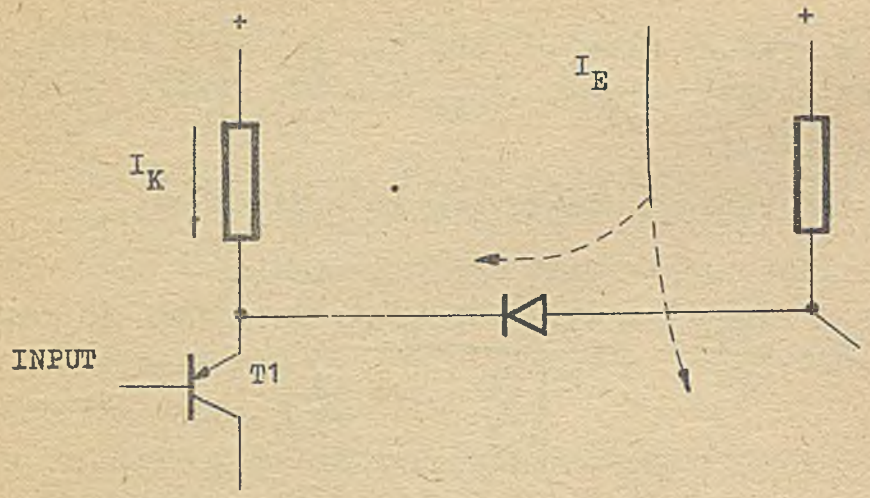


Fig. 9. Reduction of Q_E by means of the initial current I_K.

The nonlinear negative feedback can be employed instead of well-defined emitter current. This is shown in fig. 10 where, having obtained the collector current needed, the diode D1 is switched on and cuts down further increase of base current. The voltage drop across the diode D2 is introduced in order to avoid saturation.

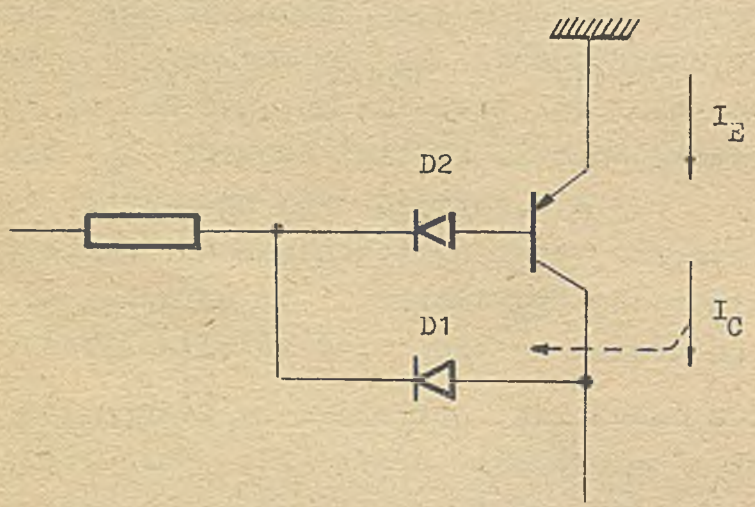


Fig. 10. The nonlinear negative feedback.

3.3. Voltage-shifting for D.C. coupling.

D.C. coupling is to be assumed because of mark-space ratio close to 0.5 /which is likely for a high-speed computer, and nonlinear operation expected. For an alternative A.C. coupling the energy stored either in the electrostatic field of the series capacitors, or in the magnetic field of the output transformers, can at least cause lack of reasonable fan-out.

However, there is one great disadvantage of D.C. coupling: the input D.C. voltage level must be restored at the output /or complemented to another allowed binary value/ and this creates the need for voltage-shifting networks. The cost is more sophisticated circuits and often much closer tolerances of voltages and of components. The usual arrangement of voltage levels can be realized for instance by means of complementary n-p-n transistors or Zener diodes [11], [13]; an interesting method is proposed by Murphy [14] who uses transistors which maintain high-speed response even for slightly positive collector-base voltage. However, these methods inconveniently increase the variety of semiconductor devices needed. Therefore, the ordinary resistive potential-divider is to be used /fig. 11/, although it also demands a relatively careful design.

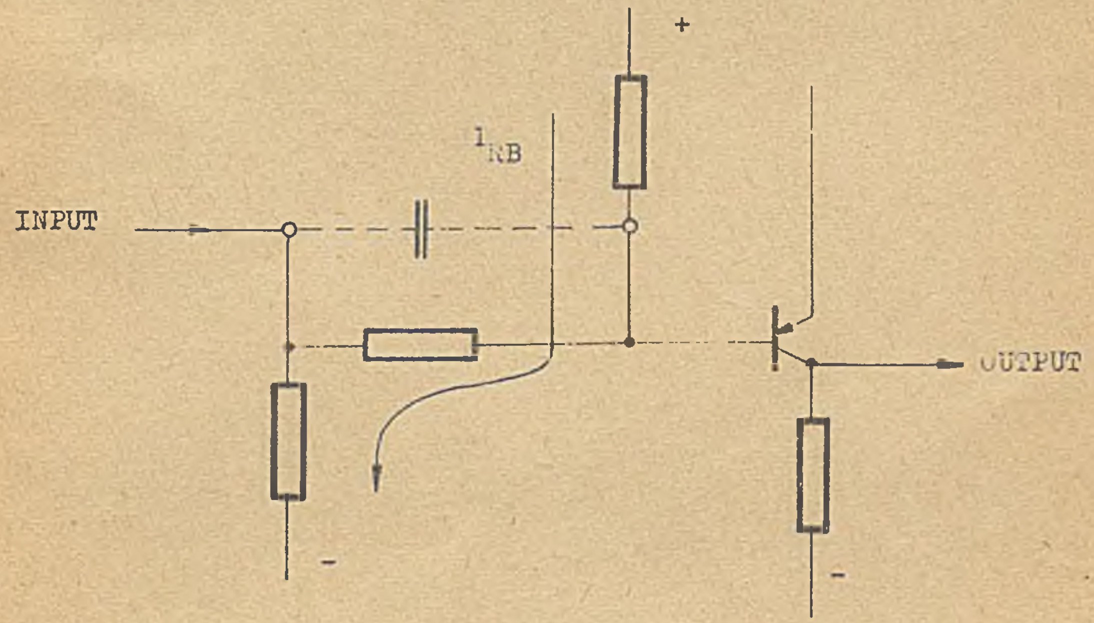


Fig. 11. The voltage-shifting principle.

It is worth to emphasize that from the point of view of voltage-shifting and tolerances, saturated circuits are the best /for further discussion refer for instance to [18]/. From the same point of view the circuits operating definitly in active region are the worst in spite of the fact that they have achieved the highest speed. Therefore, nonsaturated operation with minimum voltage-shifting provided by nonlinear negative feedback is expected by the author to be the most reasonable compromise.

APPENDIX I - The influence of Q_E.

Neglecting Q_V one obtains the formula for the required charge:

$$Q_{IN} = \int_0^t i_b(t)\,dt = Q_E + Q_B + \int_0^t \frac{Q_B}{\tau_B}\,dt$$

Or, having assumed that $\frac{Q_B}{i_c} = \tau_c$ = const, after differentiation:

$$i_b(t) = \frac{dQ_E}{dt} + \tau_c \frac{di_c}{dt} + \frac{Q_B}{\tau_B}$$

Now, putting $\frac{di_c}{dt} \cdot \frac{dQ_E}{di_c} = \frac{di_c}{dt}\tau_E$ /where $\tau_E = \frac{dQ_E}{di_c}$/ instead of $\frac{dQ_E}{dt}$, one has instead of /5/:

$$i_b(t) = \frac{di_c}{dt}\left(\tau_c + \tau_E\right) + \frac{Q_B}{\tau_B}$$

The time constant τ_E can be expressed in terms of small signal equivalent circuit /active region/ if one can put

$$Q_E = C_{Te} \cdot \Delta V_{eb'}$$

where:

$$C_{Te} = \text{const}$$

$$\frac{d\, V_{eb'}}{dI_c} = r$$

Then $\tau_E = C_{Te} \cdot r$ which is similar to the corresponding expression in [7].

APPENDIX II - "On demand" current gain β_s

Let us consider the equation describing the recombination of charge

$$\frac{dQ}{dt} + \frac{Q}{\tau_B} = i_b$$

The general solution for τ_B = const is /compare eq. /6//

$$Q(t) = \left[Q_o + \int_o^t i_b(t)\, e^{\frac{t}{\tau_B}}\, dt \right] e^{-\frac{t}{\tau_B}}$$

and in saturation $Q_o = Q(o) = I_B \tau_B = I_B \tau_{Bs}$

According to definition, "on demand" collector current is that current for which the transistor emerges from the saturation region and therefore one should substitute $\beta_0 \tau_c$ instead of the other τ_B's. Finally, at the assumed time $t = t_n$ for $i_b = I_B = \mathrm{const}$ one obtains*/

$$\beta_s = \frac{I_C}{I_B} = \beta_0 + \left[\frac{\tau_{Bs}}{\tau_c} - \beta_0\right] e^{-\frac{t_n}{\beta_0 \tau_c}}$$

References

1. SPARKES J.J., BEAUFOY R.: The Junction Transistor as a Charge Controlled Device, Proc.IRE, December 1957:45, 1740-42.
2. SPARKES J.J.: The Measurement of Transistor Transient Switching Parameters, Proc.IEE pt.B, May 1959:106, 15, 562-67.
3. BEAUFOY R.: Transistor Switching-Circuit Design Using the Charge Control Parameters, Proc.IEE pt.B, May 1959:106, 17, 1085-91.
4. KRUITHOF A.: Transient Response of Junction Transistors and Its Graphical Representation, Proc.IEE pt.B, May 1959: 106, 17,1092-1101.
5. LE CAN: Transient Behaviour and Fundamental Transistor Parameters, Electronic Appl. /Philips/, 1959-1960:20,2, 56-83.
6. SPARKES J.J.: A Study of the Charge Control Parameters of Transistors, Proc.IRE, October 1960:1696-1705.
7. DAS M.B., BOOTHROYD A.R.: Determination of Physical Parameters of Diffusion and Drift Transistors, IRE Transaction ED-8, January 1961:15-30.
8. WINDSOR R.C.: Measurement of Transistor Charge - Control Parameters by Integration of Stored Charge, ATE Journal, July 1961:17,3,95-108.
9. MOLL J.L.: Large Signal Transient Response of Junction Transistors, Proc.IRE, December 1954.

*/ $Q(t_n) = I_C \tau_c$

10. CRIPPS L.G.: Transistor High-Frequency Parameter f_1, Electronic and Radio Engineer, September 1959:341.

11. PRESSMAN A.J.: Design of Transistorized Circuits for Digital Computers, J.F.Rider Publish, New York 1959.

12. MILES R.J.: High-Speed Logic Circuits Using Diode Logical Networks and Current-Switching Amplifiers, Mullard Technical Communications, March 1961:47, 286-94.

13. MILES R.J.: Output-Coupling Networks for Use with Logical Circuits of the Emitter-Current Switching Type, Mullard Technical Communications, March 1961:47, 295-300.

14. MURPHY D.W.: High-Speed Non-Saturating Switching Circuits Using a Novel Coupling Technique, International Solid-State Conference, Digest of Technical Papers, 1962.

15. LINDMAYER J., WRIGLEY C.: The High Injection-Level Operation of Drift Transistors, Solid-State Electronics, Pergamon Press, 1961:2, 79-84.

16. BARRY J.N.: Switching Transistors Operation in the Saturation Region, Electronic Technology, June 1962:206-211.

17. SIMMONS C.D.: Hole Storage Delay Time and its Prediction, Semiconductor Products, September, October, November 1960.

18. GOLDSTICK G.H.: Comparison of Saturated and Nonsaturated Switching Circuit Techniques, IRE Trans. on El.Comp., June 1960:161-175.