THE ANALYSIS AND DESIGN OF A
FAST PULSE AMPLIFIER

by

H. Verweij
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SUMMARY

An amplifier is discussed offering a gain of \( \times 10 \) and a rise-time of 2 ns, for a driving and load impedance of 125\( \Omega \).

It accepts either a positive or a negative input signal, but delivers simultaneously a positive and a negative output pulse of maximum 7 V.

By way of a passive matched attenuator at the input, the gain can be decreased by 6, 12 or 20 db or combinations of these.

A servo-circuit, which is believed to be novel in this application, clamps the base line of the output signal, so that no d... shift occurs for high pulse rates.

Theoretical expressions are derived for the network parameters and the response.
1. INTRODUCTION

As in many other areas of electronics, the h.f. transistor has by-passed the h.f. vacuum tube in fast pulse amplifier design.

The reasons are the greater gain-bandwidth product of h.f. transistors (900 Mc/s) compared to that of normal h.f. vacuum tubes (300 Mc/s) and their smaller electrical size.

After its introduction by Percival\(^1\) and in particular after the work of Ginston et al.\(^2\), distributed amplification has commonly been used for vacuum tube wide-band amplifiers. Gain bandwidth products were obtained, which may slightly exceed that of the individual tube\(^3\).

The design of transistor distributed amplifiers is hampered by the frequency and amplitude dependency of the transistor input impedance, though possible solutions to this problem have been given\(^4-8\). They do not show any real advantage over feedback amplifiers, and only when large current swings are required in the load is it possible that distributed amplification may yield better results.

Transistor wideband feedback amplifiers have been discussed by various authors\(^9-15\).

In the described amplifier, series feedback in the emitter has been applied.

The amplifier has been developed for use in nuclear physics experiments with the CERN Proton Synchrotron and Synchro-cyclotron.

A servo-circuit has been designed, which clamps the base line of the output signal to ground level. This system limits somewhat the l.f. performance.

Variation of the gain is obtained by a passive matched attenuator at the input, thus leaving the amplifier untouched.
2. METHODS TO EXTEND THE BANDWIDTH OF A COMMON EMITTER STAGE

2.1 General

A classical way to improve the frequency response of an amplifier is by frequency dependent feedback. Minimum rise-time will be obtained when the feedback is applied over each individual stage (local feedback).

The bandwidth of a common emitter stage, the slowest of the three basic transistor configurations, can be extended by series feedback in the emitter or shunt feedback from collector to base.

Reddi\textsuperscript{13}) proves that emitter degeneration is preferable with drive from a voltage source and collector to base feedback if driven from a current source. His optimum configuration, for a low power amplifier, consists of an input transistor with emitter feedback followed by an output transistor with collector-to-base feedback, eventually cascaded.

In the described amplifier all stages have emitter feedback. The driving and output impedance happened to have such a value that in practice no difference could be observed, from the rise-time point of view, between two stages series feedback or one series and one parallel feedback.

Series feedback lends itself better for d.c. coupling and therefore better for the base-line clamping circuit. A drawback of parallel feedback is that the transistor has to deliver current into the feedback loop. These considerations led to the application of local series feedback exclusively.

2.2 The response of a common emitter stage

The equivalent circuit used in the calculations is given in Fig. 1. For an ohmic collector load the influence of $C_c$ can be translated by a capacity $C_s = (1 + A_c) C_c \approx R_t C_c / r_e$ in parallel to $C_b' e$ (Fig. 3) (see A.1). The effective gain-bandwidth product $\omega_b$ is then given by:

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\[ \omega_{t1} = \frac{\omega_t}{1 + \omega_t R_C C_c} \] (1)

where:
\[ \omega_t = \frac{1}{C_b' c r_e}. \]

For a step function input \( V_i \), the output voltage is:
\[ V_o(t) = \frac{\beta_c V_i}{R_b + r_b' c} R_c \left( 1 - e^{-t/\tau_A} \right) \] (2)

here:
\[ \tau_A = \frac{1}{\omega_{t1} \left\{ (1 - \alpha_0) + \frac{r_b}{R_b} \right\}} \] (3)
\[ R_b = r_{bb'} + R_g \]
\[ \alpha_0 \approx 1. \]

The d.c. voltage gain is:
\[ A_o = \frac{\beta_c R_c}{R_b + r_b' c} \] (4)

The voltage gain rise-time quotient:
\[ \frac{A_o}{t_r} = \frac{A_o}{2.2 \tau_A} = \frac{\omega_{t1} R_c}{2.2 R_b} = \frac{\omega_t}{2.2 \left( 1 + \omega_t R_C C_c \right)} \] (5)

2.3 The common emitter stage with feedback in the emitter

The object is to reduce the time constant of the collector current. This can be achieved by connecting a feedback resistor in the emitter.
It can easily derived that:

\[ V_o(t) = \frac{\beta_c V_i}{R_b + R_{b'e} + \beta_c R_e} \left( 1 - e^{-t/\tau_b} \right) \]  \hspace{1cm} (6)

\[ \tau_b = \frac{1}{\omega_{t2} \frac{R_b + R_e}{R_b(1 - \alpha_c) + r_e + R_e}} \]  \hspace{1cm} (7)

\[ R_e = \text{feedback resistor in emitter.} \]

\[ \omega_{t2} = \frac{\omega_t}{1 + \omega_t(R_e + R_e)C_c} \]  \hspace{1cm} (see A3) \hspace{1cm} (8)

The d.c. voltage gain:

\[ A_0 = \frac{\beta_c R_e}{R_b + R_{b'e} + \beta_c R_e} \]  \hspace{1cm} (9)

In practice \( R_e \gg R_b(1 - \alpha_c) + r_e \), so that (7) and (9) can be reduced to:

\[ \tau_b \approx \frac{1}{\omega_{t2} \frac{R_b + R_e}{R_e}} \]  \hspace{1cm} (10)

\[ A_0 \approx \frac{R_e}{R_e} \]  \hspace{1cm} (11)

The voltage-gain rise-time quotient becomes:

\[ \frac{A_0}{2.2\tau_b} = \frac{\omega_{t2}}{2.2} \frac{R_e}{R_b + R_e} = \frac{\omega_t}{2.2} \frac{R_e}{R_b + R_e} \left( 1 + \frac{1}{\omega_t(R_e + R_e)C_c} \right) \]  \hspace{1cm} (12)

Comparison of the expressions (3) and (7) shows that the rise-time has been reduced. Practical values are 5 - 10 times. However, the gain has been reduced by a greater factor, resulting in a lower voltage-gain rise-time quotient. After choosing the desired rise-time, one can find the obtainable gain by way of (12) or the reverse.
For iterative stages $R_b = R_e + r_{bb}$.

A further reduction of the rise-time can be obtained, keeping the same voltage gain, by connecting a capacity in parallel to $R_e$.

An oscillatory function is obtained for the output voltage (see A.4). For optimum response (just monotonic) the value of the parallel capacity becomes:

$$C_0 = \frac{(R_b + R_e)^2}{\frac{t}{\omega_c} R_b R_e^2}. \quad (13)$$

The expression giving the output voltage is then:

$$V_o(t) = V_1 R_e \left\{ \frac{1 - e^{-2/\tau_b t}}{R_b (1 - \omega_c) + R_e + R_e^2} + \frac{(R_e - R_b) \omega_c t^2}{R_b (R_b + R_e^2)} t e^{-2/\tau_b t} \right\}. \quad (14)$$

The rise-time of expression (14) is:

$$\tau_{r_2} = 2 \tau_b \sqrt{\frac{1}{4} - \frac{1}{32} \left( \frac{R_e + R_b}{R_b} \right)^2}. \quad (15)$$

The rise-time of (6) is in the same notation

$$\tau_{r_1} = \tau_b \sqrt{2\omega}. \quad (16)$$

A rise-time reduction factor $S$ can be defined, being:

$$S = \frac{\tau_{r_1}}{\tau_{r_2}} = \frac{1}{\sqrt{\frac{1}{2} - \frac{1}{16} \left( \frac{R_e + R_b}{R_b} \right)^2}} \quad (17)$$

In practice $S \approx 1.6.$
2.4 Shunt peaking

Another possibility of reducing the rise-time is by shunt peaking (Fig. 5). This can be applied in all cases considered.

Shunt peaking together with resistive emitter feedback gives a response function which is similar to that obtained for an R,C, network in the emitter (see A.5).

The maximum value of \( L \) still giving monotonic response is:

\[
L = \frac{1}{4} \frac{R_e + R_{b4}}{\omega t_2 (R_e + r_e)}.
\]  \hspace{1cm} (18)

\[
R_{b4} = R_c + r_{bb}.
\]

If \( R_e \gg r_e \), it can be simplified to:

\[
L = \frac{1}{4} \frac{r_b (R_e + R_{b4})}{\omega t_2}.
\]  \hspace{1cm} (19)

The output voltage is then given by:

\[
V_o(t) = IR_e \left[ \frac{R_{b4}}{R_e} \left( 1 - e^{-\frac{t}{2r_b}} \right) + \frac{R_e + r_{bb} - R_c}{R_e + r_{bb} + R_c} \right] \omega t_2 \left( e^{-\frac{t}{2r_b}} - e^{-t} \right).
\]  \hspace{1cm} (20)

and the rise-time:

\[
t_{r3} = 2r_b \sqrt{\pi \left( \frac{1}{4} - \frac{1}{32} \left( \frac{R_{b4} + R_e}{R_c} \right)^2 \right)}.
\]  \hspace{1cm} (21)

The quantity (21) is only slightly worse than that given by (15).
After having determined an optimum network for emitter peaking \((R_e)\) some improvement can still be obtained by shunt peaking. (See \(A.6\) and Fig. 6.)

It is clear that the value of \(L\) has to be smaller than given by (18).

The approximate value of the inductance is then:

\[
L_1 = \frac{R_{b1}}{R_e R_e + R_{b1}} \tau \frac{(R_e + R_{b1})}{\tau} \frac{\omega t_2}{R_e - R_{b1}}.
\]  

(22)

In practice this inductance is about half of that given by (18).

The reduction of the rise-time (15) is \(\approx 20\%\).

3. BASE LINE CLAMPING CIRCUIT

The amplifier should not introduce any base-line shift in an amplified pulse train nor show saturation effects. This might introduce timing errors or saturation in following circuitry.

As already indicated, the amplifier is a.c. coupled. The current stabilizing networks in the emitter introduce shift (Fig. 7).

For the argument the burst of pulses can be considered as one continuous pulse having the length of the burst and the amplitude

\[
V_a = V_i \frac{t_1}{t_2}.
\]

(23)

where:

\(V_i\) = amplitude of individual pulses
\(t_1\) = pulse width
\(t_2\) = \(\frac{1}{\text{pulse freq. in burst}}\)
A burst of negative pulses will move the steady emitter voltage negative by an amount:

\[ V_s = V_a \left(1 - e^{-t/\tau}\right) \]  \hspace{1cm} (24)

here:

\[ \tau = C_1 \frac{R_1 R_2}{R_1 + R_2} \approx C_1 R_2. \]

A p-n-p transistor will be cut off when:

\[ V_s > I_e R_3 + V_{be} \]  \hspace{1cm} (25)

\[ I_e = \text{quiescent emitter current} \]
\[ V_{be} = \text{quiescent base emitter voltage}. \]

By introducing another RC network in the collector, R_4 C_2, it is possible to obtain a flat burst as output, provided:

\[ \frac{R_3}{R_2} = \frac{R_4}{R_1} \]  \hspace{1cm} (26)

and

\[ C_1 R_1 = C_2 R_4. \]  \hspace{1cm} (27)

The above mentioned statements have been verified in practice. Some troubles were met when connecting to a next stage, as this load slightly upsets the compensation conditions.

The "suicide effect" (25) in the emitter makes this system less attractive. Various other ways of shift feedback have been tried, which led to the final servo-system.

Consider an amplifier with strong negative feedback (Fig. 8).

The gain of such a system will be:

\[ \frac{V_o}{V_i} = \frac{1}{1 - \Delta \alpha A_S} \]  \hspace{1cm} (28)
If $A_s A_s > 1$, this becomes:

$$\frac{V_2}{V_1} = -\frac{1}{A_s}. \quad (29)$$

When the feedback voltage is tapped off via a diode as indicated, the conditions will remain the same for negative output excursions but the positive will not be fed back.

This principle can be applied to the circuit of Fig. 7 as shown in Fig. 9. Without $C_2$ the natural shift at the collector is negative. Due to the action of the feedback amplifier this will be reduced by a factor: $1/A_s A_s$.

In actual practice the diode conducts a small amount of current ($\approx 100 \mu A$), the base current of a transistor. The feedback amplifier is a differential amplifier, which stabilizes the output d.c. level to any required value.

A positive output pulse will cut off the diode and the base current will start charging $C$. This voltage change is fed back to the output in such a way that it forces the output level back to the original d.c. level, thus differentiating the output pulse.

It is clear that the value of $C$, the base current and the loop gain determine the differentiation time constant, and so the l.f. response.

4. CIRCUIT DESCRIPTION

The amplifier consists of (Fig. 10):

a) a passive input attenuator,
b) an inverter ($T_1$)
c) an emitter follower driver ($T_2$)
d) a channel delivering the positive output ($T_3, T_7 - T_{10}$)
e) a channel delivering the negative output ($T_{12}, T_{16} - T_{19}$)
f) base-line clamping or servo-systems for both outputs, $T_4 - T_6$ and $T_{13} - T_{15}$. 

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The amplifier is completely d.c. coupled, however, the servo-circuit puts a limit to the l.f. response.

Zener diodes have been applied to get a step in d.c. level. They do not disturb the pulse response, they even gave some improvement in some cases.

Stabilization for the operating point of the individual transistor is obtained by way of high emitter resistors.

Very good d.c. stability of the complete amplifier is achieved by the action of the servo-system.

In all amplification stages local emitter feedback is used, while, where essential, shunt peaking in the collector.

Preset potentiometers adjust some d.c. levels; this method was preferred over a narrow restriction for the tolerance of the components.

The input is protected by the diodes D₃ and D₄; these limit the maximum excursion to 6 V. Protection against excessive signals at the output is obtained by D₁₁, D₂₆ and D₁₅, D₂₁.

The attenuator¹⁹) (a) gives the possibility of varying the gain in steps of 6, 12 or 20 dB and any combination of these. This method of gain adjustment was preferred over a system where the feedback of the amplifier is changed, as in that case the response is liable to change as a function of the gain. The rise-time of the attenuator is such (tᵣ < 0.5 ns) that it does not have any noticeable influence on the overall response of the amplifier. The attenuator contains matched T-pads, made up of metal film resistors and sliding switches. The impedance of the switches is ≈ 75 Ω and their feedthrough capacity very low.

A positive signal goes from the attenuator to the driver T₂ (c), a negative signal passes via the inverter T₁ (b), which has a gain of 1x, without having noticeable influence on the overall response. It has resistive feedback in the emitter, shunt peaking in the base and shift properties following (26) and (27).
The emitter follower $T_2$ (c) provides a low impedance for driving the two amplification channels. The inductances $L_8$ and $L_9$ give some series peaking and improve the loading conditions on the emitter follower. The rise-time of this part is fast compared to the final rise-time.

The first transistor ($T_3$) in the positive channel (d) gives a gain of $\approx 2.2 \times$ and a rise-time of $\approx 0.7$ ns. The theoretical value of $C_{30}$ is 10.7 pF for $f_t = 600$ Mc/s and 8.9 pF for 900 Mc/s, so rather close correspondence to the practical value (see A 7.1). Limitation of the positive output signal is arranged in this transistor. The maximum voltage which can be developed across $R_{20}$, when cutting off the current in $T_3$ is $\approx 1.6$ V. Multiplying this by the gain of the output stage ($\approx 4.5 \times$) gives 7 V. Adjustment of the overall gain ($10 \times$) is done by $R_{19}$.

The output stage consists of a double cascode. In a normal common emitter stage $\omega_t$ is reduced to $\omega_{t2}$ (8) by the feedback via $C_c$. In practice $\omega_t = (2 - 2.5) \omega_{t2}$. The cascode configuration diminishes this feedback strongly, so that $\omega_{t2}$ approximates $\omega_t$. The rise-time of the output stage is $\approx 1.8$ ns (see A 7.2).

The theoretical values of $L_7$, $C_{38}$ and $C_{39}$ are 48 nH and 14.6 pF for $f_t = 600$ Mc/s and 32 nH and 9.7 pF for 900 Mc/s. So good correspondence with practice in the case of $L_7$, while $C_{38}$, $C_{39}$ are slightly higher. This is probably because the calculations are based on no overshoot while in practice $\approx 10\%$ overshoot is accepted.

A double cascode had to be used to be capable of delivering the required output power, while also the rise-time becomes less dependent of the output amplitude.

Connecting $T_7$, $T_9$ and $T_8$, $T_{10}$ in a distributed manner gave no reduction in rise-time.
The input transistor, $T_{12}$, in the negative output channel (e) gives the same gain and rise-time as $T_3$ in the positive channel. It also limits the output signal and the adjustment of the gain is possible with $R_{21}$.

Transistor $T_2$ had to be added in order to create the correct polarity for the n.p.n. output transistors.

A cascode has also been applied here for the reasons explained above. A single transistor can be used at the top, as the power-handling capability of this transistor is sufficiently high ($300 \text{ mW l.s.o.}, 30 \text{ mW for the p.n.p.}$).

The base-line clamping circuit (f') for the positive output channel is formed by the diode $D_{13}$, the differential amplifier $T_5$ and $T_6$, and the emitter follower $T_4$, which feeds the output d.c. level back into the amplification chain.

The diode $D_{13}$ passes the base current of $T_5$, so for d.c. the loop is closed. The differential amplifier stabilizes the output d.c. level to that set by $R_{71}$, being ground level. Very stable d.c. conditions for the amplifier are obtained this way. To improve the temperature stability $R_9$ has been added.

The natural shift of the output, for a pulse train, is negative and so this will be strongly reduced by the feedback. In theory by a factor $1/A_0 A_3 \approx 1/120$, in practice the residual shift can be neglected (see Fig. 13).

A positive pulse will cut off $D_{13}$ and the base current of $T_6$ will charge $C_{34}$ at a rate of $\approx 1.5 \mu\text{V/ns}$. The ramp will be amplified by the differential amplifier. The emitter follower $T_4$, driven by this ramp will charge $C_{31}/C_{32}$ via $R_{31}$. And only the voltage across $C_{31}/C_{32}$ will be fed back to the output and start reducing the output signal. The introduction of $C_{31}/C_{32}$ and $R_{21}$ delays the instant at which the ramp starts having noticeable influence on the output signal. The time constant $C_{32} R_{21}$ has to be short compared to the shift-creating time constants in the amplifier not to disturb the clamping action and it is obvious that it governs the l.f. response.

The absolute value of the reduction is independent of the amplitude of the output signal and will be most important for small signal amplitudes.
A flat pulse-train output means constant current operating point of the output transistors. The average current increases during a burst, which will discharge $C_{37}$, $C_{41}$. The servo-system corrects the base voltage of $T_7$, $T_9$ so that the current operating point remains constant. To limit the base excursion the emitter impedance had to be kept low, which led to the adjustable voltage source $T_{11}$.

The diodes $D_{10}$ and $D_9$ are included for protection.

The clamping action in the negative output channel is performed by $D_{22}$, $T_{13}$, $T_{14}$ and $T_{15}$.

5. **PERFORMANCE**

Many of the described amplifiers are in use in CERN experiments and their behaviour has shown to be reliable and stable over long periods.

They are constructed as a plug-in element, the wiring being mounted on a special card, which has a conducting layer (copper) on one side. The classical care has to be taken when building the amplifier, such as short leads, non-inductive decouplings, etc.

The rise and fall-times of a typical amplifier are shown in Fig. 11. The test-wave form is generated by a mercury-wetted contact relay pulse generator ($t_r \approx 0.3$ ns) and displayed on a sampling oscilloscope (with improved mixer, $t_r \approx 0.45$ ns). The spread in rise-time of the produced amplifiers is from 1.6-2.2 ns for the positive output and from 2.0-2.5 ns for the negative output. Figure 12 gives rise and fall-times as a function of the output voltage. The augmentation of the rise-time for high output voltages is mainly due to increase of $f_e$.

Figure 13 illustrates the amplification of a burst of pulses, 200 ns long, at a rate of 1/3 c/s, consisting of 25 ns wide pulses spaced by 100 ns ($2 \times 10^6$ pulses) and simulating the conditions in a beam of the CERN Proton Synchrotron. The remaining base-line shift is seen to be very small.
The maximum width of an individual pulse which can be handled is mainly determined by the time constants in the feedback loop. A droop of 10% occurs after $\approx 5 \mu\text{sec}$ at all output levels and an initial output signal of 100 mV will be reduced to $\approx 30 \mu\text{V}$ after $\approx 50 \mu\text{sec}$, an output of 4 V is brought down to $\approx 1.5 \text{mV}$ after $\approx 250 \mu\text{sec}$.

Repetitive pulses will be amplified normally. The average output current from the positive output should not exceed 6 mA and from the negative output 25 mA, in order not to damage the output transistors.

The amplifier linearity is represented by Fig. 14, for both positive and negative output. These curves have been measured not going via the inverter. The performance of the inverter is indicated separately.

The output noise level is 75 $\mu\text{V}$ when the output is matched.

The temperature dependency of the rise-time, gain and d.c. levels, has been examined over the range from 10-50°C. The rise-time of the positive output increased from 1.8-2.3 ns (2.0 at 25°C) and that of the negative from 2.2-2.4 ns (2.3 at 25°C). The gain increased by 3%. The output d.c. level of the positive output goes positive at a rate of $\approx 0.5 \text{mV/°C}$ and that of the negative output goes negative by the same amount.

The delay of a positive signal from the input to the positive output is 6.5 ns and to the negative output 7.5 ns. A negative input signal is delayed by 7.5 ns and 8.5 respectively.

6. ACKNOWLEDGEMENTS

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APPENDIX

1) Influence of $C_o$

The analysis will be executed starting from the equivalent circuit \(^{(16)}\) of Fig. 1.

The feedback via the "Miller" capacity $C_o$, in the simple common-emitter circuit, has been accurately calculated by Koss \(^{(17)}\) for a collector load consisting of a capacity and resistance in parallel. It is transformed by a resistance $R_s$ and capacity $C_s$ in series and this in parallel to $C_{b'o}$ (Fig. 2).

\[
R_s = \frac{C_o}{g_m C_o} \frac{1}{1 + \frac{\omega^2 C_o^2}{g_m^2}}
\]  
\hspace{1cm} (1A)

\[
C_s = g_m C_o R_c \frac{1 + \frac{\omega^2 C_o^2}{g_m^2}}{1 + \frac{\omega^2 C_o^2 R_s}{g_m}}
\]  
\hspace{1cm} (2A)

For a very small capacitive load and not too high a frequency the expressions may be simplified to:

\[
R_s = \frac{C_o}{g_m C_o}
\]  
\hspace{1cm} (3A)

\[
C_s = g_m C_o R_c
\]  
\hspace{1cm} (4A)

In the case of a pure ohmic load, so $C_t = 0$, $R_s$ vanishes. In the following only $C_s$ will be considered, also for small capacitive loads. The results thus obtained are slightly pessimistic as $R_s$ weakens the influence of $C_s$. 

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A new current gain-bandwidth product can be defined which includes the feedback via \( C_c \):

\[
\omega_{t_1} = \frac{\omega_t}{1 + \omega_t R_c C_c}
\]

(5A)

here: \( \omega_t = 1/C_{b_e} r_{e^*} \).

2) The common-emitter stage

The response of the common-emitter circuit (Fig. 3) for a step function input \( V_i \) is given by (2).

For current drive \( R_b >> r_e \), then

\[
\tau_a \approx \frac{1}{\omega_{t_1} (1 - \alpha_o)} ;
\]

(6A)

while for voltage drive \( R_b = r_{bb^*} \), and

\[
\tau_a = \frac{1}{\omega_{t_1} \left( 1 - \alpha_o + \frac{r_e}{r_{bb^*}} \right)} .
\]

(7A)

3) Change of \( \omega_t \) due to feedback in the emitter

The current gain-bandwidth product \( \omega_{t_2} \) in this case becomes:

\[
\omega_{t_2} = \frac{\omega_t}{1 + \omega_t (R_c + R_e) C_c}
\]

(8A)

This has been derived following the system used by Ekiss\(^7\), Bruun\(^9\) and Readi\(^13\) take \( \omega_{t_1} = \omega_{t_2} \).
Measurements on the applied transistors 2N 769, and 2N 743 confirmed (8A) however. In practical cases the theoretical difference between $\omega_{t_1}$ and $\omega_{t_2}$ is of the order of 20%.

4) An R.C. network in the emitter

The equivalent circuit of Fig. 4 is used. For a step function input $V_1^*$, using the Laplace notation, one can write:

$$\frac{V_1}{p} = i_b \frac{R_b}{R_b + \frac{r_e}{(1 - \alpha_0) + \frac{1}{\omega_{t_2}^2}}} + (1 + \beta) i_b \frac{R_e}{1 + \frac{1}{p} \frac{C_e}{R_e}}$$  \hspace{1cm} (9A)

$$i_b(p) = \frac{V_1}{p} \frac{1}{R_b + \frac{r_e}{(1 - \alpha_0) + \frac{1}{\omega_{t_2}^2}}} + (1 + \beta) \frac{R_e}{1 + \frac{1}{p} \frac{C_e}{R_e}}$$  \hspace{1cm} (10A)

say: $C_e R_e = r_e$,

and: $\beta = \frac{\beta_o}{1 + \beta_o \frac{1}{\omega_{t_2}^2}}$; where: $\beta_o = \frac{\alpha_o}{1 - \alpha_o}$.

Taking $\omega_{t_2}$ here is somewhat severe, a value between $\omega_{t_2}$ and $\omega_{t_1}$, is probably more realistic.

Further: $i_c(p) = \beta i_b$.
\[ i_c(p) = \frac{V_i}{p} \frac{\beta_o}{1 + \beta_o \frac{p}{\omega_{t_2}}} \frac{1}{R_b + \frac{r_e}{(1 - \alpha_o) + \frac{p}{\omega_{t_2}}} + \left(1 + \frac{1}{1 + \beta_o \frac{p}{\omega_{t_2}}} \right) \frac{R_e}{1 + p r_e}}. \]

(11A)

After some rearrangement:

\[ i_c(p) = \frac{V_i}{p} \frac{1}{R_b \left(1 - \alpha_o + \frac{p}{\omega_{t_2}}\right) + \frac{1}{1 + p r_e} \frac{r_e + R_e + p (r_e + R_e)}{\omega_{t_2}}}. \]

(12A)

The assumption has been made that \( R_e \gg R_b (1 - \alpha_o) \). A further development gives:

\[ V_c(p) = \frac{V_i}{p} \frac{R_c}{R_b} \frac{1 + p r_e}{\frac{r_e}{\omega_{t_2}} R_b \left\{ \frac{R_b (1 - \alpha_o) + r_e + R_e}{\omega_{t_2}} \right\} + \frac{R_b}{\omega_{t_2}} + R_b (1 - \alpha_o) \frac{R_e + r_e + p^2}{\omega_{t_2}} + \frac{r_e}{\omega_{t_2}} R_b}. \]

(13A)

This function represents a damped oscillation. Choosing the two poles real and equal gives the situation of critical compensation, the response will still be monotonic then.

From this condition the required emitter time constant can be derived:

\[ \frac{R_b}{\omega_{t_2}} + R_b (1 - \alpha_o) \frac{r_e + R_e + p r_e}{\omega_{t_2}} \left(\frac{r_e}{R_b}\right)^2 - 4 \frac{R_b (1 - \alpha_o) + r_e + R_e}{\omega_{t_2}} \frac{r_e}{R_b} = 0 \]

(14A)

\[ \{R_b + R_e + R_b (1 - \alpha_o) \omega_{t_2} r_e + \omega_{t_2} r_e R_b \}^2 - 4 \omega_{t_2} r_e R_b \{R_b (1 - \alpha_o) + r_e + R_e\} = 0. \]

(15A)
In general:

\[ \omega_{t_2} \tau_e \left\{ r_e + R_b (1 - \alpha_0) \right\} \ll R_e + R_b. \]  

(16A)

Applying this in (15A) gives:

\[ (R_b + R_e)^2 - 4 \omega_{t_2} \tau_e R_b \left\{ R_b (1 - \alpha_0) + r_e + R_e \right\} = 0 \]  

(17A)

Solving for \( \tau_e \) yields:

\[ \tau_e = \frac{(R_b + R_e)^2}{4 \omega_{t_2} R_b \left\{ R_b (1 - \alpha_0) + r_e + R_e \right\}}. \]  

(18A)

Simplified:

\[ \tau_e = \frac{(R_b + R_e)}{4 \omega_{t_2} R_b R_e}. \]  

(19A)

and

\[ C_e = \frac{(R_b + R_e)^2}{4 \omega_{t_2} R_b R_e^2}. \]  

(20A)

or in other form:

\[ \tau_e = \frac{R_b + R_e}{4 R_b} \tau_b. \]  

(21A)

The poles of (13A) will lie at:

\[ \rho_{1,2} = -\frac{R_b}{\omega_{t_2}} \frac{R_b (1 - \alpha_0) \tau_e + R_e}{2 \tau_e R_b} + \frac{R_e + \tau_e R_e}{2 \tau_e R_b}. \]  

(22A)

Using the simplification (16A):

\[ \rho_{1,2} = -\frac{R_b + R_e}{2 \tau_e R_b} = -\frac{2}{\tau_b}. \]  

(23A)
The inverse transform for (13A) gives:

\[ V_o(t) = V_i \frac{\omega_{ts}}{\tau_e R_b} \left[ \frac{1}{p^2} \left( 1 - e^{pt} \right) + t e^{pt} \frac{1}{p + \tau_e} \right]. \] (24A)

After putting in the value for \( p \) and some rearrangements:

\[ V_o(t) = V_i R_e \left[ \frac{1 - e^{-2/\tau_b t}}{R_b (1 - \alpha_e) + R_e + R_e} + \frac{(R_e - R_b) \omega_{ts} t e^{-2/\tau_b t}}{R_b (R_e + R_e)} \right]. \] (25A)

The rise-time of expression (25A) can be found using Elmore's \(^{17}\) definition of rise-time for a network with monotonic response. This differs about 10\% from the commonly used 10-90\% value, but is much simpler to compute.

Putting the derived value for \( p \) in (13A) gives:

\[ V_o(p) = \frac{V_i}{p} \frac{\omega_{ts}}{\tau_e R_b} \frac{R_e}{(1 + p \tau_e)} \left( \frac{1 + p \tau_e}{1 + p \tau_b + p^2 \frac{\tau_b}{1 + p \tau_b}} \right). \] (26A)

The rise-time of this function, being the transform of the output signal, is:

\[ t_{r_2} = 2\tau_b \sqrt{\pi \left[ \frac{1}{1 + \tau_b} \left( \frac{R_e + R_b}{R_b} \right)^2 \right]}. \] (27A)

In the case of the emitter resistance without a parallel capacity the rise-time is:

\[ t_{r_1} = \tau_b \sqrt{2\pi}. \] (28A)

(In the 10-90\% notation \( t_r = 2.2 RC \).)
A rise-time reduction factor can be defined:

$$S = \frac{t_{r1}}{t_{r2}} = \sqrt{\frac{1}{2} - \frac{1}{16} \left( \frac{R_e + R_b}{R_b} \right)^2}.$$  \hspace{1cm} (29A)

In practice $S \approx 1.6$.

5) A shunt peaking inductance in the collector (Fig. 5)

Assume a current step $I_1$, which normally is the collector current of a preceding transistor.

The relation exists:

$$\frac{I_1}{\beta} (R_{\beta 1} + pL) = I_0 \left( R_{\beta 1} + r_{bb'} + pL \right) + \frac{r_e}{(1 - \alpha_0) + p} + (1 + \beta) R_e$$  \hspace{1cm} (30A)

$$i_b(p) = \frac{I_1}{\beta} \frac{R_{\beta 1} + pL}{R_{\beta 1} + r_{bb'} + pL + \frac{r_e}{(1 - \alpha_0) + p} + (1 + \beta) R_e}$$

$$\beta = \frac{\beta_0}{1 + \beta_0 \frac{p}{\omega_t}}, \quad R_{b1} = R_{\beta 1} + r_{bb'}, \quad i_0 = \beta i_b,$$

$$i_c(p) = \frac{I_1}{\beta} \frac{R_{\beta 1} + pL}{\frac{R_{b1} + R_e}{\beta_0} + R_e + r_e + p \left( \frac{L}{\beta_0} + \frac{R_e + R_{b1}}{\omega_t^2} \right) + p^2 \frac{L}{\omega_t^2}}$$  \hspace{1cm} (31A)
A simplification can be made by stating:

\[
\frac{R_e + R_b}{\beta_0} \ll R_e + r_e \quad \text{and} \quad \frac{L}{\beta_0} \ll \frac{R_e + R_b}{\omega t_2}
\]

\[
i_c(p) = \frac{I_i}{p} \frac{R_c + pL}{R_e + r_e + p\left(\frac{R_e + R_b}{\omega t_2}\right) + p^2 \frac{L}{\omega t_2}} \quad (32A)
\]

Assuming again that the best response will be obtained for equal poles, then \( L \) can be derived.

\[
\left(\frac{R_e + R_b}{\omega t_2}\right)^2 - 4 \frac{L}{\omega t_2} (R_e + r_e) = 0 \quad (33A)
\]

\[
L = \frac{1}{4} \frac{(R_e + R_b)^2}{\omega t_2 (R_e + r_e)} \quad (34A)
\]

If \( R_e \gg r_e \):

\[
L = \frac{1}{4} r_b \left( R_{b_1} + R_e \right) \quad (35A)
\]

The poles of \( (32A) \) are:

\[
p_{1,2} = -\frac{2\omega t_2}{R_e + R_{b_1}} R_e = -\frac{2}{r_b}
\]

\[
(36A)
\]

which are equal to those found in the case of a parallel capacity in the emitter.

The inverse transform for \( (32A) \) gives:
\[ I_c(t) = I_1 \left[ \frac{R_e}{R_b} \left( 1 - e^{-t/\tau_b} \right) + \left( \frac{R_e + \tau_{bb}'}{R_e + \tau_{bb}'} \right) t e^{-t/\tau_b} \right] \]

(37A)

The rise-time of (37A) is:

\[ t_{r3} = 2 \tau_b \sqrt{\frac{1}{4} - \frac{1}{32} \left( \frac{R_e + R_{c1}}{R_{c1}} \right)^2} \]

(38A)

When comparing (27A) and (38A), it is clear that the parallel capacity gives a slightly better result, the difference being negligible.

6) Combination of emitter peaking and shunt peaking

The collector current \( I_C \) is expressed by (13A), in which \( R_b \) has been replaced by \( R_{b1} + pL \) and \( V_1 = I_1 (R_{c1} + pL) \), thus giving:

\[ I_C(p) = \frac{I_1}{p} \left\{ \frac{(R_{c1} + pL)(1 + p \tau_e)}{R_{b1} + pL(1 - \alpha_0) + R_e + \frac{R_{b1} + pL \tau_e}{\omega t_2} + \frac{R_e \tau_e}{\omega t_2}} + \frac{R_e \tau_e}{\omega t_2} \right\} \]

(39A)

The aim is to calculate values for \( L \) and \( \tau_e \), which give critical compensation. For this, only the denominator will be considered.
Assume:

\[(R_{b_1} + pL)(1 - \alpha_0) + r_e + R_e \approx R_e\]

\[\frac{R_{b_1} + pL}{\omega_{t2}} + (R_{b_1} + pL)(1 - \alpha_0) \frac{R_e}{\omega_{t2}} + \frac{R_e}{\omega_{t2}} + \tau_e r_e \approx \frac{R_{b_1} + pL + R_e}{\omega_{t2}}\]

Remains of the denominator:

\[\frac{R_e \omega_{t2}}{\tau_e} + p \frac{R_{b_1} + R_e}{\tau_e} + p^2 (R_{b_1} + \frac{L}{\tau_e}) + p^3 L\]  \hspace{1cm} (40A)

From the condition, equal negative poles to give optimum monotonic response, a value for L can be solved equal to:

\[L_4 = \frac{R_{b_1} \tau_{e_4} (R_e + R_{b_4})}{9 R_e \tau_{e_4} \omega_{t2} - R_e - R_{b_4}}\]  \hspace{1cm} (41A)

In order to make all poles coincide it is also necessary to change \(\tau_e\) to:

\[\tau_{e_1} = \frac{(R_{b_1} + R_e)[3(R_{b_1} + R_e) - \sqrt{3(R_{b_1} + R_e)^2 - 3R_e(R_{b_1} + R_e)}]}{9 \omega_{t2} R_e \{R_{b_1} + 3R_e - \sqrt{3(R_{b_1} + R_e)^2 - 3R_e(R_{b_1} + R_e)}\}}\]  \hspace{1cm} (42A)

The multiple pole is:

\[p_{1,2,3} = a = -\frac{1}{3} \left(\frac{R_{b_1}}{L_4} + \frac{1}{\tau_{e_1}}\right)\]  \hspace{1cm} (43A)

Now Eq. (39A) can be reduced to:

\[i_o(p) = \frac{I_1 \omega_{t2}}{p \tau_e L} \frac{R_{e_1} + p(\tau_e R_{e_1} + L) + p^2 \tau_e}{(p + a)^3}\]  \hspace{1cm} (44A)
The inverse transform of this function is:

\[ i_o(t) = I_1 \left( \frac{\omega t_2}{\tau_{e1}} \right) \frac{-R_{e1}}{L_1} + e^{-at} \left[ t^2 \left( \frac{R_{e1} R_{e1} + L_1}{2} - \frac{a L_1 \tau_{e1}}{2a} - \frac{R_{e1}}{2a} \right) \right. \]

\[ + t \left( L_1 \tau_{e1} - \frac{R_{e1}}{a^2} - \frac{R_{e1}}{a^3} \right) \]  \hspace{1cm} (45A)

The rise-time of (45A) is:

\[ t_{r1} = \frac{2}{\pi} \sqrt{\frac{L_1 \tau_{e1}}{2 R_{b1} + R_{e}} - \frac{L_1 + \tau_{e1}}{2 R_{e}}} \] \hspace{1cm} (46A)

The expression (42A) will only give real values for \( \tau_e \) if \( 3 R_e > R_{b1} \). In the described amplifier this is not the case (see A7.2).

If in expression (41A) the value for \( \tau_e \) as given by (19A) is used an error is made, but the obtained value (see A7.2) gives good correspondence to the practical optimum.

7) Calculation of the rise-time

As an example, the rise-time of the positive output channel will be calculated.

The inverter \( T_1 \) has negligible influence on the amplifiers' rise-time. The rise-time loss in the input attenuator and the emitter follower \( T_2 \) can also be neglected, remains \( T_3 \) and the output cascode (\( T_{7-10} \)).

The parameters of the 2N769 are supposed to be \(^{10} \):

\[ R_{bb'} = 25 \, \Omega \]
\[ C_0 = 1.5 \, \text{pF} \]

The minimum gain bandwidth product is 600 Me/s and the typical value is 900 Me/s.
The rise-time will be examined for these two values of \( f_t \). In effect \( f_t \) is non-linear, but is assumed to be constant over the range of interest.

I. The response of \( T_2 \)

The positive influence of \( L_3 \) and the negative effect of the load of \( T_1,2 \) on \( T_2 \) are neglected.

Transistor \( T_3 \) is driven from emitter follower \( T_2 \) which is supposed to have an ohmic output impedance, \( R_o = 10 \, \Omega \).

Further: \( R_o = 50 \, \Omega \) and \( R_e = 120 \, \Omega \).

For \( f_t = 600 \, \text{Mc/s} \)

The effective gain bandwidth product (6A):

\[
\omega_{t2} = \frac{\omega_t}{1 + \omega_t (R_e + R_o) C_e} \leq \frac{\omega_t}{1 + 2\pi \times 6 \times 10^8 (120 + 50) 15 \times 10^{-13}} \leq \frac{\omega_t}{1.96}
\]

The desired value of \( C_e \) is (20A):

\[
C_e = \frac{(R_e + R_o)^2}{4 \omega_{t2} R_e R_o}
\]

\[
C_e = \frac{(35 + 50)^2}{4 \times \frac{2\pi \times 6 \times 10^8}{1.96} \times 35 \times 50^2} = 10.7 \, \text{pF}.
\]

The rise-time is given by (27A):

\[
t_{r2} = 2 \frac{R_e + R_o}{R_e \omega_{t2}} \sqrt{\pi \left( \frac{1}{4} - \frac{1}{32} \left( \frac{R_o + R_e}{R_e} \right)^2 \right)}
\]

\[
t_{r2} = 2 \frac{35 + 50}{50 \times \frac{2\pi \times 6 \times 10^8}{1.96}} \sqrt{\pi \left( \frac{1}{4} - \frac{1}{32} \left( \frac{50 + 35}{35} \right)^2 \right)}
\]

\[
= 0.80 \, \text{ns}.
\]
The familiar 10-90% value is $0.9 \times 0.80 = 0.72$ ns.
This is the rise-time of the current pulse driving the collector load of $T_3$.

For $f_c = 900$ Mc/s.

\[
\omega_{t_2} = \frac{\omega_t}{1 + 2\pi \times 9 \times 10^8 \times 1.2 \times 10^{-13}}
\]

\[
= \frac{\omega_t}{2.44}.
\]

The required value of $C_e$:

\[
C_e = \frac{(35 + 50)^2}{4 \cdot 2\pi \times 9 \times 10^8 \times 2.44 \times 35.50^2} = 8.9 \text{ pF}.
\]

The rise-time:

\[
t_{r_2} = 2 \frac{(35 + 50)}{50 \times 2\pi \times 9 \times 10^8 \times 2.44} \sqrt{\pi \left(\frac{1}{4} - \frac{1}{32} \left(\frac{50 + 35}{35}\right)\right)}
\]

\[
= 0.67 \text{ ns}.
\]

The 10-90% value: $0.9 \times 0.67 = 0.60$ ns.

In practice 10 pF is used for $C_e$.

II. The response of $T_7 - T_{10}$.

The two parallel cascodes, $T_7-T_8$ and $T_9-T_{10}$ are driven from the same source impedance ($R_s$). This complicates the calculation. It can be shown that the derived expressions are still valid if $R_g$ is replaced by $2R_s$ in the formula for $t_r$ (27A) and $C_e(20A)$. 

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The expression for the peaking inductance $L$ becomes:

$$L_1 = \frac{1}{2} \frac{R_{b_2}}{9 R_e \tau_{e_2}} \frac{R_e + R_{b_2}}{\omega_t - R_e - R_{b_2}}$$

(42A)

in which:

$$R_{b_2} = 2 R_{t_1} + r_{bb}'$$

$r_{e_2}$ is that value of $r_e$ in which $R_b$ has been replaced by $R_{b_2}$.

The cascode circuit has been applied to reduce $C_e$ feedback. A small amount is still present due to $R_e$ and the inductive input impedance of the upper transistor ($T_b$). Measurements confirmed that for this configuration $\omega_{t_2} \approx \omega_t$.

The rise-time loss in $T_b$ is neglected (grounded base).

$$R_e = 40 \ \Omega \quad R_{b_2} = 2 R_{t_1} + r_{bb}' = 265 \ \Omega$$

For $f_t = 600 \ \text{Mc/s}$.

The desired value of $C_e$:

$$C_e = \frac{(R_{b_2} + R_e)^2}{4 \omega_t R_{b_2} R_e^2}$$

$$= \frac{(2 \cdot 120 + 25 + 40)^2}{4 \cdot 2 \pi \cdot 6 \cdot 10^9 \cdot 265 \cdot 40^2}$$

$$= 14.6 \ \text{pF}.$$ 

The rise-time, without considering $L$, is:
\[ t_{r2} = 2 \frac{R_{b2} + R_e}{R_e \omega_t} \sqrt{\pi \left( \frac{1}{4} - \frac{1}{32} \left( \frac{R_e}{R_{b2}} \right)^2 \right)} \]

\[ = 2 \frac{120 + 25 + 40}{40 \times 2\pi \times 6 \times 10^8} \sqrt{\pi \left( \frac{1}{4} - \frac{1}{32} \left( \frac{40 + 265}{265} \right)^2 \right)} \]

\[ = 3.28 \text{ ns.} \]

The 10-90\% value, \( 0.9 \times 3.28 = 2.95 \text{ ns.} \)

The shunt peaking inductance:

\[ L = \frac{1}{2} \frac{265 \times 40 \times 146 \times 10^{-13} (40 + 265)}{9 \times 40 \times 40 \times 146 \times 10^{-13} \times 2\pi \times 6 \times 10^8 - 40 - 265} \]

\[ = 48 \text{ nH.} \]

The rise-time reduction due to \( L \) is \( \approx 20\% \), taking this into account gives:

\[ t_r = 0.8 \times 2.95 = 2.34 \text{ ns.} \]

For \( f_t = 900 \text{ Mc/s.} \)

\[ C_e = \frac{6}{9} \times 14.6 = 9.7 \text{ pF.} \]

The rise-time without \( L \):

\[ t_{r2} = \frac{6}{9} \times 2.95 = 1.97 \text{ ns.} \]

and the shunt peaking inductance:

\[ L_1 = \frac{1}{2} \frac{265 \times 40 \times 97 \times 10^{-13} \times 305}{9 \times 40 \times 40 \times 97 \times 10^{-13} \times 2\pi \times 9 \times 10^8 - 40 - 265} \]

\[ = 32 \text{ nH.} \]
The final rise-time:

\[ t_r = 0.8 \times 1.97 = 1.58 \text{ ns}. \]

In practice, \( C_e = 16 \text{ pF} \), so somewhat higher than calculated. This is because the calculations are based on non-overshoot conditions, while in practice \( \approx 10\% \) overshoot is obtained and accepted. In some cases \( C_e \) had to be reduced. The applied value of \( L \) ranges from 34-52 nH so very good agreement with the theoretical value.

III. The rise-time of the amplifier

The resultant rise-time of the amplifier is:

\[ t_{r_a} = \sqrt{\sum t_r^2} \]

For \( f_t = 600 \text{ Mc/s} \)

\[ t_{r_a} = \sqrt{(0.72)^2 + (2.34)^2} = 2.44 \text{ ns}. \]

For \( f_t = 900 \text{ Mc/s} \)

\[ t_{r_a} = \sqrt{(0.60)^2 + (1.58)^2} = 1.68 \text{ ns}. \]

In practice the rise-time ranges from 1.6-2.2 ns so somewhat more favourable, this is due to the fact that \( C_e \) has been taken somewhat higher than the theoretical value.
Fig. 1 - Equivalent circuit

Fig. 2 - $C_c$ transformed into $C_s$ and $R_s$

$C_s = g_m C_c R_l$ and $R_s = \frac{C_l}{g_m C_c}$
Fig. 3 - Collector load purely ohmic. $C_1 = 0$ so $R_s = 0$

Fig. 4 - Feedback in the emitter
Fig. 5 - Shunt peaking

Fig. 6 - Emitter and shunt peaking
Fig. 7 - Shift compensation by equal time constants in emitter and collector
Fig. 8 – An amplifier with unipolar feedback

Fig. 9 – Shift compensation by servo system
fig. 11. Rise and fall time of amplifier, pos. and neg. output.
$V_o = 2.5\, V$
- a and d sweep speed 5 ns/div
- b, c, e and f " 1 ns/div.
Fig 12 - Rise and fall time vs output voltage

a) Pos. output

b) Neg. output
fig. 13. Amplification of a pulse train

hor. sens  50ms/div.
vert. sens  2V/div.

a  Input
b  Pos. output
c  Neg. output
% gain \( V_0 = 2.5V \)

- inv.
- pos.
- neg.

Fig. 14 - Linearity of gain
fig. 15. Views of wiring
Fig. 16  View of front panel