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RSIC-583

DIFFERENTIATION CIRCUITS FOR AMPLITUDE-MODULATED SIGNAL ENVELOPES

by
V. I. Anisimov
Yu. F. Likhodiyevskiy

Priborostroenie, 7, No. 5, 41-46 (1964)

Translated from the Russian
by R. C. Taylor

N67 12230

(ACCESSION NUMBER)	(THRU) 1
(PAGES) 9	(CODE)
TMX-57999 (NASA CR OR TMX OR AD NUMBER)	09 (CATEGORY)

August 1966

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Two differentiation circuits for amplitude-modulated signal envelopes in which the functions of the modulator and the demodulator are accomplished by one switch are examined in this article.

A serious problem arising during the designing of a system with signal modulation is the creation of economic and stable compensation circuits. There are two basic methods of compensation circuits which operate on alternating current.

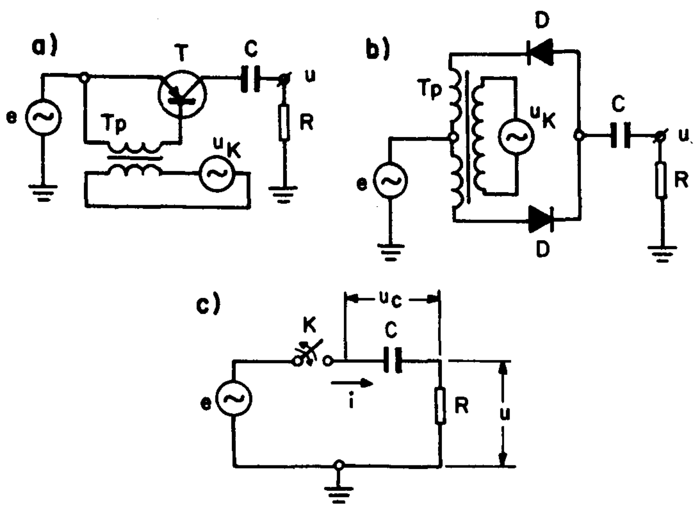

Figure 1. Schematics of Differentiating Circuits (a, b) and Their Equivalent Schematic (c)

The first method is compensation by sections of alternating current which are different resonance circuits tuned to the carrier frequency. On the strength of this such circuits are extremely critical to changes of the carrier frequency and magnitudes of the self-capacitances, inductances, and resistances.

The second method is compensation with double conversion of the signal frequency (according to the schematic: demodulator - compensation circuit of direct current - modulator). However, such devices are complex and unwieldy and also not very effective in case the construction of a compensation circuit of the differentiating type is required since it still contains filters of lower frequencies for feeding pulses to the circuit's output.

A development of the second method of compensation is the application of a circuit in which one switching element executes simultaneously the functions of the demodulator and the modulator.[1,2]

In this case the compensation circuits are significantly simplified, and during the construction of differentiating (leading) type circuits there is no necessity for using the smoothing filter.

Two simple schematics of compensation circuits of the differentiating type, in which the switching element simultaneously executes the functions of the demodulator and the modulator, are presented in this article. These schematics are presented in Figure 1a and 1b.

If we assume the transistor to be the ideal switch key, then the first of the schematics which is presented can be exchanged for the equivalent schematic (Figure 1c). If we fulfill the conditions $U_{Km} \gg E_m$, where U_{Km} is the voltage of the switching voltage and E_m is the amplitude of the incoming signal, we can also relate Figure 1b to this schematic.

In the equivalent schematic (Figure 1c) the switch key closes in step and in phase with the incoming signal.

We will find the voltage at the capacitor $u_{c\Delta}$ and the output voltage u during the feeding of sinusoidal voltage with frequency ω_0 into the input.

We will assume that

$$RC \gg \frac{1}{\omega_0} ,$$

(ω_0 is the circular carrier frequency), since the voltage at the capacitor u_c barely changes during the semiperiod of switching and is equal to the mean value U_c.

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[1] E. V. Bohn, Demodulator Lead Networks. IRE Trans. Circuit Theory, v. CT-7, 1960, N 1.

[2] E. V. Bohn, A Simple Method for the Analysis of Demodulator Compensating Networks. IRE Trans. Circuit Theory, v. CT-9, 1961, N 3.

Then in the semiperiod which corresponds to the closed switch key the current through the capacitor is

$$i_c = \frac{e - U_c}{R} \quad .$$

In the following semiperiod $i_c = 0$.

The mean current through the capacitor for the entire period is

$$I_c = \frac{1}{2\pi}\int_0^{2\pi} i_c(\omega_0 t)\, d\,(\omega_0 t) = \frac{1}{2\pi}\int_0^{\pi} \frac{E_m}{R} \sin \omega_0 t\, d\,(\omega_0 t) -$$

$$- \frac{1}{2\pi}\int_0^{\pi} \frac{U_c}{R}\, d\,(\omega_0 t) = \frac{E_m}{\pi R} - \frac{U_c}{2R} \ .$$

Considering that $C \dfrac{dUc}{dt} = I_c$, we have

$$T \cdot \frac{dU_c}{dt} + U_c = \frac{2}{\pi} E_m \ ,$$

where $T = 2RC$.

Consequently,

$$U_c = \frac{2}{\pi} E_m \left(1 - e^{-\frac{t}{T}}\right)$$

The output voltage in the operating semiperiod is

$u = e - U_c = E_m \sin \omega_0 t - U_c$.

In the second semiperiod $u = 0$ (Figure 2).

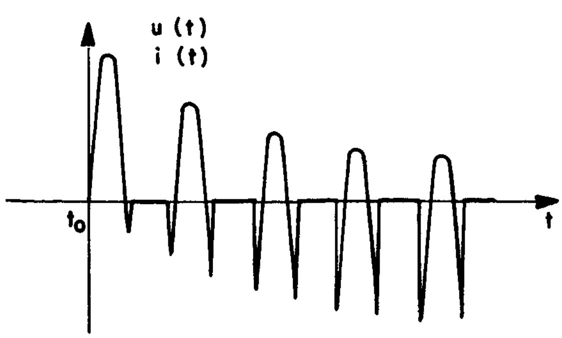

Figure 2. The Output Voltage of the Differentiating Circuit

The amplitude of the first harmonic of the output voltage is

$$U_{m1} = \frac{1}{\pi}\int_0^{\pi} u\,(\omega_0 t)\,\sin \omega_0 t\; d\,(\omega_0 t) =$$

$$= \frac{1}{\pi}\int_0^{\pi} (E_m \sin \omega_0 t - U_c)\,\sin \omega_0 t\; d\,(\omega_0 t) =$$

$$= \frac{1}{2} E_m - \frac{2}{\pi} U_c = E_m \left[\left(\frac{1}{2} - \frac{4}{\pi^2}\right) + \frac{4}{\pi^2}\, e^{-\frac{t}{T}}\right];$$

$$U_{m1} \approx \left(0.1 + 0.4 e^{-\frac{t}{T}}\right) E_m \,.$$

Apparently there are other harmonic components in the output signal besides the basic harmonic.

We will evaluate their magnitude.

The constant component is

$$U_0 = \frac{1}{2\pi}\int_0^{\pi} u\,(\omega_0 t)\; d\,(\omega_0 t) = \frac{E_m}{\pi} - \frac{U_c}{2} = \frac{1}{\pi}\, e^{-\frac{t}{T}}\, E_m.$$

The quadratic component of the first harmonic is

$$U_{m1} = \frac{1}{\pi}\int_0^{\pi} u\,(\omega_0 t)\,\cos \omega_0 t\; d\,(\omega_0 t) = 0 \,,$$

i. e., the device does not permit a phase displacement on the carrier frequency.

The sine component of the second harmonic is

$$U_{m2} = \frac{1}{\pi}\int_0^{\pi} u\,(\omega_0 t)\,\sin 2\,\omega_0 t\; d\,(\omega_0 t) = 0 \,.$$

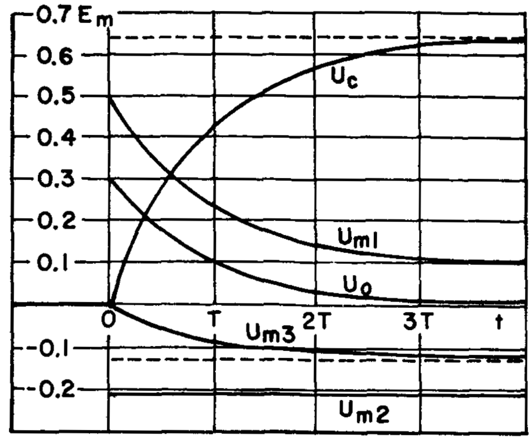

Figure 3. The Comparative Magnitude of the Harmonic Components of the Output Voltage of the Differentiating Circuit

The cosine component of the second harmonic is

$$U'_{m2} = -\frac{2}{3\pi} E_m .$$

The sine component of the third harmonic is

$$U_{m3} = -\frac{2}{3\pi} U_c = \frac{4}{3\pi^2}\left(e^{-\frac{t}{T}} - 1\right) .$$

The cosine component of the third harmonic is

$$U'_{m3} = 0 .$$

The comparative magnitude of the harmonics in time is shown in Figure 3.

Inasmuch as the presence of a demodulator which basically reacts only to the first harmonic of voltage U_{m1} is assumed in the channel of the amplifier channel, then the action of the constant component and the higher harmonics only leads to a decrease in the dynamic band of the amplifier channel.

Writing the expression for the amplitude of the first harmonic of the output signal in the form

$$U_{m1} = A \cdot a \left(1 + \frac{1 - a}{a} e^{-\frac{t}{T}}\right) E_m ,$$

where

$$A = \frac{U_{m1}(0)}{E_m} \approx 0.5 ;$$

$$a = \frac{U_{m1}(\infty)}{U_{m1}(0)} \approx 0.2 ,$$

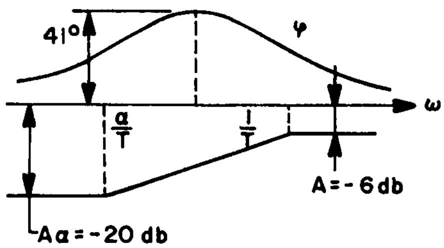

Figure 4. The Equivalent Frequencies of the Characteristic of a Differentiating Circuit

it is not hard to be convinced that such a transient characteristic matches the equivalent transfer function along the signal envelope.

$$W_{or}(p) = \left|\frac{U_{m1}(p)}{E_m(p)}\right| = A \cdot a \cdot \frac{\frac{T}{a} p + 1}{Tp + 1} ,$$

where T = 2RC.

Thus, the given circuit is actually a differentiating circuit for an amplitude-modulated signal envelope.

The corresponding frequency characteristics are depicted in Figure 4.

As has already been noted, the quoted conclusion is valid at $T \gg (1/\omega_0)$. The nearness of the carrier frequency to 1/T decreases the effectiveness of the circuit (coefficient A drops and α approaches 1)

During practical realization of the schematic it is necessary to consider the resistance introduced by the switch key and the internal resistance of the signal source. The effect of the resistance of the source R_i and the resistance of the key r consists of an increase of the constant time T and a decrease of coefficient A, i.e.,

$$T = 2 (R + R_i + r) C;$$

$$A = 0.5 \cdot \frac{R}{R + R_i + r} .$$

Since magnitude α and, consequently, the maximum phase shift do not depend on values R_i and r, the circuits being considered can easily match the different amplifying cascades. In particular, we can build a differentiating circuit of the second order if we bypass two similar devices with the aid of a transistor cascade (Figure 5).

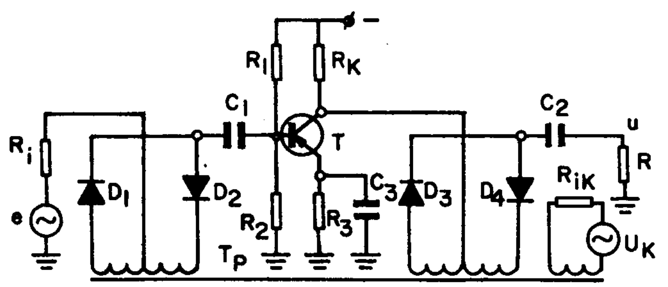

Figure 5. A Differentiating Circuit of the Second Order.

Thus, combining the functions of the modulator and the demodulator in one switching element permits schematics for differentiation along amplitude-modulated signal envelopes to be built which do not contain smoothing filters.

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3. REPORT TITLE

DIFFERENTIATION CIRCUITS FOR AMPLITUDE-MODULATED SIGNAL ENVELOPES

Priborostroenie, 7, No. 5, 41-46 (1964)

4. DESCRIPTIVE NOTES *(Type of report and inclusive dates)*

Translated from the Russian by R. C. Taylor

5. AUTHOR(S) *(Last name, first name, initial)*

Anisimov, V.I. and Likhodiyevskiy, Yu. F.

6. REPORT DATE	7a. TOTAL NO. OF PAGES	7b. NO. OF REFS
16 August 1966	10	2

8a. CONTRACT OR GRANT NO. N/A b. PROJECT NO. N/A c. d.	9a. ORIGINATOR'S REPORT NUMBER(S) RSIC-583 9b. OTHER REPORT NO(S) (Any other numbers that may be assigned this report) AD________

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