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By

K. -H. Schmelovsky, and H. Boerner

EDITED TRANSLATION

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OPTIMUM DEMODULATOR FOR AMPLITUDE-MODULATED FM-SUBCARRIERS

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Report from the Central Institute for Solar and Terrestrial Physics (Heinrich-Hertz-Institute) of the German Academy of Sciences, Berlin

Submitted 10 Nov 1971.

According to the optimization process for Markov input processes [1] introduced by Stratonovich solutions have already been given for simple AM-, PM-, and FM-demodulators [2], [3]. The following article deals with a complicated type of demodulator, the amplitude-modulated frequency-modulation subcarrier. The modulation process is assumed to be a Markov process and the interference as an additive, white Gaussian process.

1. Statement of the Problem

From the input process

(1) $$y(t) = s(t) + n(t) = A_0 \cos\left\{\omega_0 t + \int_{-\infty}^{t} \Delta\omega [1 + \eta(\sigma)] \times \cos\omega_s \sigma \, d\sigma\right\} + n(t)$$

[A_0 amplitude, ω_0 carrier center frequency, ω_s subcarrier frequency, $1 + \eta$ modulation process, $\Delta\omega$ frequency deviation, n normal white noise, $\bar{n} = 0$, $\overline{nn_\tau} = (N_0/2)\,\delta(\tau)$] η(t) is to be determined with a minimum root-mean-square error. A priori η(t) should obey the stochastic equation

(2) $$\dot{\eta}(t) = -\beta\eta(t) + \xi(t)$$

[$\bar{\xi} = 0$, $\overline{\xi\xi_\tau} = \chi\delta(\tau)$]. Thus η(t) is assumed to be an RC-limited, normal stochastic process with the limiting frequency β, $\bar{\eta} = 0$, and the dispersion $\sigma_\eta^2 = \chi/2\beta$. With η(t) the second summand also changes stochastically in the phase of the signal $\omega_0 t + \Phi$. According to Eq. (1)

(3) $$\dot{\Phi} = \Delta\omega [1 + \eta(t)] \cos\omega_s t$$

The equation system (2), (3) describes a two-dimensional vector-Markov process with the components Φ, η. After reception of the input process y(t) the signal parameters Φ, η are not known exactly due to the superimposed disturbance n(t), but rather can only be described statistically by the a posteriori probability density $w(\eta, \Phi/y)$. It results according to Bayes' formula and its temporal change can be described according to Stratonovich [1] by an expanded Fokker-Planck equation which, for the problem being dealt with here, has the form

(4)
$$\frac{\partial w(\Phi, \eta, t)}{\partial t} = -\frac{\partial}{\partial \Phi} \Delta\omega (1+\eta) w \cos \omega_s t + \beta \frac{\partial}{\partial \eta} \eta w + \frac{\chi}{2} \frac{\partial^2 w}{\partial \eta^2} + (F - F) w$$

$[F = -(y-s)^2/N_0]$. The first two summands on the right side take the drift into account and the third, the diffusion of the probability distribution which result from the a-priori data according to Eq. (2) and (3) and the last summand takes into account the information obtained through the reception of the input signal y(t).

The solution of the nonlinear partial differential equation (4) yields the a-posteriori probability distribution whose maximum with respect to η represents the sought optimum estimated value of the modulation parameter $\hat{\eta}$, while the dispersion square σ_η^2 of the a-posteriori distribution gives the error dispersion of the evaluation.

3. Design of the Optimum System

If we limit ourselves to the case of a small a-posteriori dispersion we can use a two-dimensional Gaussian statement for the approximate solution of Eq. (4)

(5) $$w(\Phi, \eta) = \exp\left\{c - \frac{1}{2|M|}\left[(M_{\Phi\Phi})(\Phi - \hat{\Phi})^2 - 2(M_{\Phi\eta})(\Phi - \hat{\Phi})(\eta - \hat{\eta}) + (M_{\eta\eta})(\eta - \hat{\eta})^2\right]\right\}$$

(c constant, $|M|$ determinant of the correlation matrix M of Φ, η; $(M_{\alpha\beta})$ adjuncts of the element $k_{\alpha\beta}$ of M). If we substitute it, then a coefficient comparison yields the following system of equations for the velocity coordinates of the mean values of $\hat{\Phi}$, $\hat{\eta}$ and $w(\Phi, \eta)$

(6) $$\dot{\hat{\Phi}} = \Delta\omega\,(1 + \hat{\eta})\cos\omega_s t - k_{\Phi\Phi}\frac{2A_0}{N_0}\,y(t)\sin(\omega_0 t + \hat{\Phi})$$

(7) $$\dot{\hat{\eta}} = -\beta\hat{\eta} - k_{\Phi\eta}\frac{2A_0}{N_0}\,y(t)\sin(\omega_0 t + \hat{\Phi})$$

(8) $$\dot{k}_{\Phi\Phi} = 2\Delta\omega\,k_{\Phi\eta}\cos\omega_s t - k_{\Phi\Phi}^2\frac{2A_0}{N_0}\,y(t)\cos(\omega_0 t + \hat{\Phi})$$

(9) $$\dot{k}_{\Phi\eta} = \Delta\omega\, k_{\eta\eta} \cos \omega_s t - \beta\, k_{\Phi\eta} - k_{\Phi\Phi}\, k_{\Phi\eta} \frac{2 A_0}{N_0} \cos(\omega_0 t + \hat{\Phi})$$

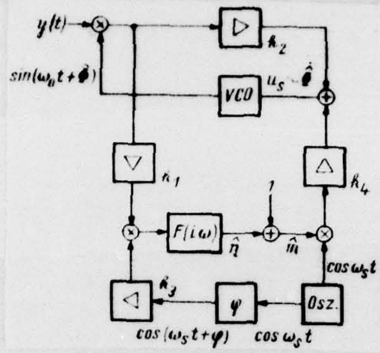
(10) $$\dot{k}_{\eta\eta} = -2\beta k_{\eta\eta} + \chi - k_{\Phi\eta}^2 \frac{2 A_0}{N_0} y(t) \cos(\omega_0 t + \hat{\Phi})$$

This system of equations supplies the sought algorithm of the demodulator system. Equations (6) and (7) are realized by an analog circuit. The nonlinear coefficients $k_{\Phi\Phi}$ and $k_{\Phi\eta}$ are determined from the nonlinear sytem of equations (8), (9), (10). A solution of this system is found through the statement

(11) $$k_{\Phi\Phi} = \text{const.} = k_{\Phi 0};\ k_{\Phi\eta} = k_{\Phi\eta 0} \cdot \cos(\omega_s t + \varphi); \\ k_{\eta\eta} = \text{const.} = k_{\eta 0}$$

With this statement the equation system (6) and (7) is realized by the circuit according to Fig. 1.

Fig. 1. Fundamental circuit of the demodulator.



It consists of a phase-lock base circuit of the first order with a voltage-controlled oscillator (VCO) and an auxiliary circuit which is between the phase-demodulator output and the summing member in front of the VCO. The auxiliary circuit contains an oscillator oscillating with the subcarrier frequency. The subcarrier is amplitude-modulated with the process $1 + \hat{\eta}$ which is first acquired from the reception process by coherence demodulation and filtered in $F = \beta/(\beta + i\omega)$. The optimum frequency control results additively both through the modulated local subcarrier and also through the error voltage on the phase detector output of the base circuit. The gains of the four amplifier stages are

(12) $$k_1 = -\frac{2A_0}{N_0\beta}; \quad k_2 = -\frac{2A_0}{N_0}k_{\varphi 0}; \quad k_3 = k_{\varphi\eta 0}; \quad k_4 = \Delta\omega$$

The estimated value of the modulation process $1 + \hat{\eta}(t)$ can be removed in the auxiliary circuit following addition of 1 behind the filter F. One achieves a smaller demodulation error if one obtains the output process through amplitude modulation of the control voltage of the VCO $u_s \sim \dot{\phi}$ [cf. Eq. (3)]. The values of $k_{\phi 0}, k_{\eta 0}, k_{\phi\eta 0}$ and ϕ were determined in the following manner: from equations (8), (9), (10), and (11) one obtains the system

(13) $$0 = \Delta\omega\, k_{\phi\eta 0} \cos\varphi - k_{\phi 0}^2 \cdot K$$

(14) $$-\omega_s k_{\phi\eta 0} \sin(\omega_s t + \varphi) = \Delta\omega\, k_{\eta 0} \cos\omega_s t - \beta k_{\phi\eta 0} \times \cos(\omega_s t + \varphi) - K k_{\phi 0} k_{\phi\eta 0} \cos(\omega_s t + \varphi)$$

(15) $$0 = 2\beta(\sigma_\eta^2 - k_{\eta 0}) - \frac{K \cdot k_{\phi\eta 0}^2}{2}$$

whereby $K = A_0^2/N_0$ is the temporal mean value of $-(2A_0/N_0)$ [y cos $(\omega_0 t + \hat{\phi})$ assuming a small estimate error $|\phi - \hat{\phi}| \ll 1$.

In this system of equations one can first express $\varphi, k_{\phi\eta 0}$ and $k_{\eta 0}$ by $k_{\phi 0}$

(16) $$\tan\varphi = -\frac{\omega_s}{\beta + K k_{\phi 0}}$$

(17) $$k_{\phi\eta 0} \cos\varphi = \frac{K}{\Delta\omega} k_{\phi 0}^2$$

(18) $$k_{\eta 0} = \sigma_\eta^2 - \frac{K^2 k_{\Phi 0}^4}{4\beta \Delta\omega^2 \cos^2\varphi}$$

Therewith one obtains an algebraic equation for determination of $k_{\Phi 0}$.

With the substitutions

(19) $$y = \frac{K}{\beta} k_{\Phi 0}$$

(20) $$s = 1 + \frac{\omega_s^2}{\beta^2}$$

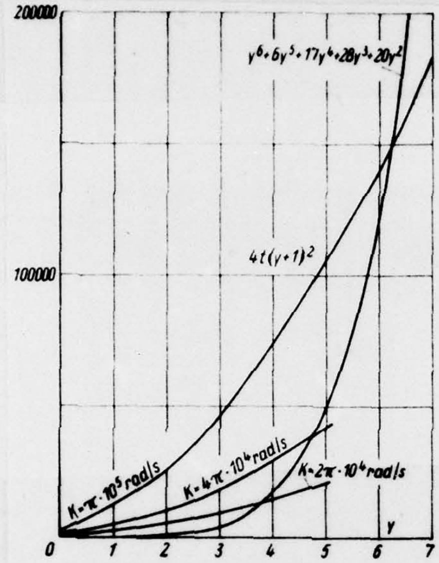
(21) $$t = \sigma_\eta^2 \Delta\omega^2 \frac{K}{\beta^3}$$

it acquires the form

(22) $$y^6 + 6y^5 + (12 + s)\, y^4 + 4\,(2 + s)\, y^3 + 4sy^2 = 4t\,(y + 1)^2$$

This equation was solved for the applied case of a weather satellite receiver (APT-system). The following numerical values were used for this: $\sigma_\eta^2 = 1$, $\Delta\omega = 3.14 \cdot 10^4$ rad/s, $\beta = 0.75 \cdot 10^4$ rad/s, $\omega_s = 1.5 \cdot 10^4$ rad/s. The solutions for various values of K (signal-noise ratio at the demodulator input) can be read from the diagram in Fig. 2 as abscissas of the intersecting points of both sides of the curves corresponding to Eq. (22), and in the given case, further approximated analytically. Therewith the sought coefficients and gains of the realization circuit are known from Fig. 1.

Fig. 2. Graphic solution of Eq. (22).



4. Properties of the System.

Fig. 3 shows the dependence of the dispersion of the phase error in the demodulation circuit on the signal-to-noise ratio for the studied case of application. For comparison a curve is drawn in for the dispersion results from the quasilinear treatment of the problem using the Wiener-Hopf equation. According to the values of practical experience [4] [5] [6] the phase-lock circuit is stable roughly in the range $\sigma_\varphi^2 \leq 0.273$ in which the curves are drawn through and unstable with larger σ_φ^2 values (dotted curves). Thus the demodulation threshold is determined. One realizes that it is reduced about 4.5 dB in comparison with the quasilinear optimization.

The 3-dB bandwith of the phase-lock base circuit of the demodulation circuit

$$B_0 = V_g = \frac{A_0}{2} k_2 = \frac{A_0^2}{N_0} k_{\Phi 0} \quad (23)$$

which is equal to the total amplification of the circuit is dependent on the signal-to-noise ratio as shown in Fig. 4. With weak interferences the filter also picks up the weaker components of the modulation spectrum which lie further from the carrier so that the estimation accuracy increases. With a stronger interference on the other hand there results a narrower bandwidth. By frequency selection one aims for the possible optimum of the estimation accuracy.

Fig. 3. Dispersion of the phase-demodulator error depending on the signal-to-noise ratio at the demodulator input. KEY: 1) quasilinear method; 2) nonlinear method.

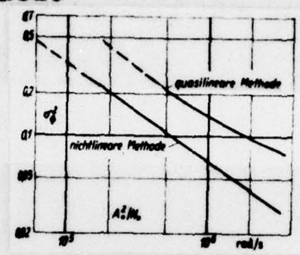
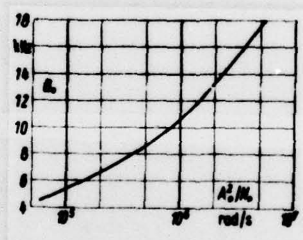


Fig. 4. Bandwidth of the base circuit depending on the signal-to-noise ratio.



The auxiliary circuit in contrast to the base circuit has a supplementary amplification V_H, so that for the total amplification V = $V_g(1+V_H)$. One obtains

(24) $$V_H = \frac{k_{\Phi 0} A_0^2}{2\beta N_0}$$

Fig. 5 shows the dependence of this supplementary amplification on the signal-to-noise ratio. Under the fundamental reception conditions the supplementary amplification in the auxiliary circuit is about three times.

The auxiliary circuit processes the preliminary knowledge about the signal structure by controlling the VCO with the subcarrier frequency. The controlled oscillator thus carries out the subcarrier oscillation from the very beginning. It is passively readjusted only to its amplitude changes. This is possible (see Fig. 3) with considerably higher accuracy than in the case of a modulator without an auxiliary circuit.

Fig. 5. Supplementary amplification of the auxiliary circuit depending on the signal-to-noise ratio.

Summary

Based on the theory of Markov processes according to the nonlinear method an optimum demodulator was developed for amplitude-modulated FM-subcarriers [7]. A phase-lock circuit resulted which consists of a base circuit and an auxiliary circuit. The auxiliary circuit contains a generator for the subcarrier frequency. Such a circuit was first used in the the weather satellite receiving system WES 2 which was developed in the Heinrich-Hertz Institute and resulted in a considerable reduction of the demodulation threshold.

The practical difference between non-optimum demodulation in a simple phase-lock circuit and optimum demodulation can be seen in Fig. 6. Reception of a weather picture was switched between the demodulators every 30 seconds. The noise interferences with non-optimum demodulation are particularly clear at the edge of the picture.

Fig. 6. Comparison of optimum and non-optimum demodulation (Sahara is in the middle of the photo).

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