(12) **Oversettelse av europeisk patentskrift**

(11) **NO/EP 2786096 B1**

NORGE

(19) NO
(51) Int Cl.

G01C 19/5776 (2012.01)

Patentstyret

(21)	Oversettelse publisert	2016.06.06
(80)	Dato for Den Europeiske Patentmyndighets publisering av det meddelte patentet	2016.01.20
(86)	Europeisk søknadsnr	12780654.5
(86)	Europeisk innleveringsdag	2012.10.15
(87)	Den europeiske søknadens Publiseringsdato	2014.10.08
(30)	Prioritet	2011.12.01, DE, 102011119949
(84)	Utpekte stater	AL AT BE BG CH CY CZ DE DK EE ES FI FR GB GR HR HU IE IS IT LI LT LU LV MC MK MT NL NO PL PT RO RS SE SI SK SM TR
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(54)	Benevnelse	**REGULATING DEVICE, ROTATION RATE SENSOR AND METHOD FOR OPERATING A REGULATING DEVICE WITH HARMONIC DESIRED VALUE SIGNAL**
(56)	Anførte publikasjoner	EP-A1- 2 336 717

Description

The invention refers to a control device, in particular to a rotation rate sensor device with harmonic command variables or harmonic set point signals. The invention further refers to a method for operating a control device with harmonic command variables.

Conventional control methods are tailored to control problems with constant or only slowly changing command variables, wherein the value of a controlled process variable affected from a disturbance is kept as close as possible to a predetermined set point, or respectively is updated as close as possible to a changing set point. Some applications as for example micromechanical rotation rate sensors for analysis of a Coriolis force provide an excitation of an oscillator with its resonance frequency and with a defined oscillation amplitude by a control loop. In this process, a controller controls the force signal generated by it such that the difference between a predetermined harmonic set point signal and a measured oscillator movement vanishes.

In this process, typically a measurement signal which reflects the movement of an oscillator along a direction of excitation is at first fed to a demodulator. The demodulator multiplies the measurement signal with a harmonic signal with a angular frequency ωd, which corresponds to the resonance angular frequency ω0 of the oscillator. The control itself is performed with a set point signal in the baseband being constant or in any case independent from the resonance angular frequency. The output signal of the controller gets then re-modulated in a modulator onto a harmonic signal with an angular frequency ωm, which corresponds to the resonance frequency ω0 of the oscillator. The modulation product is then compared with the predetermined set signal. The difference between the two signals controls finally an actuator, which performs based on the controller signal a force to the oscillator such that the oscillator oscillates according to the predetermined set oscillation. As the control is performed in the baseband, a low pass filter filters the frequency conversion products subsequent to the modulator, in particular at the double resonance frequency, by which process, however, the bandwidth of the controller and hence its reaction speed to changes of the deflection is limited.

Document EP 2336717 discloses the preamble of the independent claims.

At the application date not yet published German patent application DE 102010055631.9 a control with a harmonic set point signal in the frequency band of the resonance angular frequency ω0 of the oscillator is described.

In many fields of application the oscillator oscillates after deactivating of the drive with a decaying amplitude. Is the control activated during a decay phase of the oscillator the activation time depends on the phase and amplitude difference between the decaying oscillation and the set point signal.

Object of the invention is a control concept for improving the switch-on behaviour of a control device for a harmonic command variable. The object is solved by the subject-matter of the independent claims. Further embodiments are given in the corresponding dependent claims.

In the following embodiments of the invention, their functioning as well as their advantages will be described based on the Figures. Elements of the embodiments may be combined with each other, insofar they do not exclude each other.

Figure 1 illustrates a schematic block diagram of a device with a control device according to an embodiment of the invention, which comprises a controller main unit for controlling a harmonic oscillation based on a harmonic set point signal and a controller extension unit for synchronizing the harmonic set point signal.

Figure 2A illustrates a simplified block diagram with details of the controller extension unit of Figure 1 according to another embodiment.

Figure 2B illustrates a block diagram with further details of the controller extension unit of Figure 2A according to another embodiment.

Figure 3A illustrates a schematic block diagram of a device with a control device according to an embodiment, which refers to a controller main unit with a continuous PI controller for harmonic set point signals and a dead time element.

Figure 3B illustrates schematically the transfer function of the PI controller according to Figure 3A.

Figure 4A illustrates a schematic block diagram of a device with a control device according to an embodiment which refers to a controller main unit with a discrete PI controller for harmonic set point signals and a dead time element.

Figure 4B illustrates schematically the transfer function of the controller main unit of Figure 4A.

Figure 5A illustrates a schematic block diagram of a device with a control device according to an embodiment, which refers to a controller main unit with a discrete PI controller for harmonic set point signals and a controller extension working similarly to band pass.

Figure 5B illustrates schematically the transfer function of the controller extension of Figure 5A.

Figure 6A is a schematic top view of the micromechanical part of a rotation rate sensor according to another embodiment of the invention.

Figure 6B is a schematic cross-sectional view of the micromechanical part of the rotation rate sensor of Figure 6A.

Figure 6C is a schematic block diagram of the rotation rate sensor according to Figures 6A and 6B.

Figure 7 is a schematic top view of the micromechanical part of a rotation rate sensor according to another embodiment of the invention.

Figure 8 illustrates a simplified process flow for a method for operating a control device.

The device 100 as shown in Figure 1 comprises an oscillator 190 and a control device with a controller main unit 200 and a controller extension unit 600. The oscillator 190 is a mass, which is moveably suspended along a direction of excitation and is capable to oscillate along the direction of excitation with a resonance frequency $\omega 0$. In the stationary case the oscillator 190 performs a translational or rotational oscillation of the resonance angular frequency $\omega 0$. According to an embodiment the oscillator 190 is an excitation unit, a Coriolis unit or a detection unit of a rotation rate sensor. The rotation rate sensor may be for example formed as an MEMS (microelectromechanical system).

The sensor 170 captures the movements of the oscillator 190 and outputs a measurement signal, which reflects the whole deflection of the oscillator 190 along a direction of excitation. The measurement signal corresponds to a controller input signal for the controller main unit 200. The controller main unit 200 compares the controller input signal with a harmonic set point signal output from the controller extension unit 600 and generates based on the signal difference a controller output signal, which is output to an actuator unit 180. The controller main unit 200 determines the controller output signal such that the actuator unit 180 generates a force signal such that the difference between the predetermined harmonic set signal and the measured movement of the oscillator 190 vanishes.

According to an embodiment the controller extension unit 600 comprises an actuator activation unit 695 by which the actuator unit 180 can be activated. In the block diagram of Figure 1 the actuator activation unit 695 is illustrated as a switch in the supply of the controller output signal to the actuator unit 180, wherein the switch is controllable by the output signal of the controller extension unit 600.

The controller extension unit 600 is activated by the control device, for example by switching on an operation voltage. The actuator unit 180 stays deactivated at first. Depending on the previous history, the oscillator 190 is addressed or performs a residual oscillation if the actuator unit 180 is deactivated. The controller extension unit 600 determines while the actuator unit 180 is still deactivated from the measurement signal an actual phase and an actual amplitude of such a residual oscillation of the oscillator 190 and supplies a synchronized set point signal adapted to the actual phase and the actual amplitude to the controller main unit 200. As soon as the harmonic set point signal or the synchronized control signal deduced therefrom is available, the controller extension unit 600 activates the actuator unit 180 via the actuator activation unit 695 such that the amplitude of the residual oscillation of the oscillator 190 is phase synchronously amplified to the set amplitude and such that in this process the energy contained in the residual oscillation is used.

The controller extension unit 600 allows to set the oscillator starting from all considerable initial states in very short time to its set amplitude and to maintain it there. A residual oscillation occurs in particular then if the oscillator 190 is to be brought to oscillate with the set amplitude after deactivating the drive or the actuator unit 180 at an arbitrary time within the decay time.

According to an embodiment the control devices 200, 600 and the oscillator 190 are constituents of a rotation rate sensor of a navigation instrument, in particular of a navigation instrument for an aircraft which has to be transferred into an undisturbed operation state after a short time power breakdown in a time as short as possible. The controller extension unit 600 estimates amplitude and phasing of the present decaying residual oscillation of the oscillator 190.

According to an embodiment the controller extension unit 600 comprises a Kalman filter for estimating the amplitude and phasing. From the amplitude and phasing of the residual oscillation an initial phase as well as an initial amplitude for a phase and amplitude correct switch-on of the harmonic set point signal are deduced under consideration of internal signal runtimes. The present residual oscillation is used such that the period of recomissioning after switch-off for a short time is reduced. The larger the amplitude of the present residual oscillation, the less time is necessary in order to let the oscillator oscillate with the set amplitude again. Hence, in particular after short breakdowns of the power supply recommissioning time (ramp up time) of a system comprising the control device 200, 600 is reduced.

According to the embodimen illustrated in Figure 1 the controller extension unit 600 comprises a capturing unit 610 and a synchronization unit 620. The capturing unit 610 determines from the measurement signal the actual phase and the actual amplitude of the residual oscillation of the oscillator 190 at least in an activation phase of the device 100, for example after switching on the operation voltage again. From the actual phase and the actual amplitude as well as further system parameters, e.g. signal runtimes and signal retardation times, the capturing unit 610 determines synchronisation information, which indicates the phase and amplitude of the harmonic set point signal generated by the synchronisation unit 620. The synchronisation unit 620 receives the synchronisation information and transfers the harmonic set point signal determined by the synchronisation information to the controller main unit 200. For example, the controller main unit 200 comprises a summation unit 221 which forms from the harmonic set point signal output from the synchronisation unit 620 and the measurement signal a difference signal.

According to an embodiment the resonance angular frequency ωr of the harmonic set point signal is predetermined by the resonance angular frequency ω0 of the oscillator 190, wherein this frequency is also integrated as initial value in the estimation for the actual phase and the actual amplitude. According to another embodiment the controller extension unit 600 comprises a temperature capturing unit, wherein the capturing unit 610 bases the estimation of the actual phase and actual amplitude on a temperature corrected resonance angular frequency of the oscillator 190, and wherein the harmonic set point signal oscillates with the temperature corrected resonance angular frequency.

According to the embodiment illustrated in Figure 1 the controller extension unit 600 comprises a frequency storage unit 630, which stores frequency information describing the current oscillation frequency of the oscillator 190 in temporal intervals. According to an embodiment the frequency storage unit 630 stores the current resonance angular frequency periodically in a non-volatile storage. For example, the update frequency is chosen such that application typical temperature changes may be followed. According to an embodiment the update frequency is in a range from 1 Hz to 100 Hz, for example around 10 Hz.

The controller extension unit 600 retrieves the frequency information for estimating the actual phase and the actual amplitude of the residual oscillation of the oscillator 190 and/or uses the frequency information stored there for generating the harmonic set point signal, for example for controlling of an oscillator generating the set point signal. In the relevant time periods for recommissioning, this means within the decay time of the oscillation of the oscillator 190, the temperature and therefore the resonance angular frequency of the oscillator 190 hardly changes such that the value for the driving frequency stored last during ongoing operation in a non-volatile storage represents after a restart a sufficiently goody approximate value for the actual resonance angular frequency of the oscillator 190 and may be used as the initial value for the control. For example, oscillators in micromechanical rotation rate sensors have time constants in a range of 10 s. After deactivating the force transmission, for example after loss of an operation voltage, the oscillator oscillates after about 30 s still with about 5 % of the set amplitude.

According to an embodiment a sequence control of the capturing unit 610 controls the actuator activation unit 695 such that the actuator unit 180 is only activated then if the controller extension unit 600 outputs a phase and amplitude synchronous harmonic set point signal. According to an embodiment the actuator activation unit 695 is a switching device, for example a digital switch, which supplies the actuator unit 180 with the controller output signal of the controller main unit 200 only then if a phase and amplitude correct harmonic set point signal is available to the controller main unit 200.

According to another embodiment the capturing unit 610 deactivates at least after determining the synchronisation information such partial units of the controller extension unit 600 which are not needed anymore.

Figure 2A illustrates details of the capturing unit 610. According to an embodiment the capturing unit 610 comprises a filter unit 612. For example, an analogue measurement signal is sampled with a sampling time T and converted into a digital measurement signal in this process. The filtering unit 612 estimates from the sampling values of the measurement signal, from an estimation value for the variance of a measurement noise contained in the measurement signal, and from an estimation value for a constant amplitude offset of the measurement signal estimation values for a variation in time of the residual oscillation, for example the estimated zero points. According to an embodiment the filter unit 612 uses in this process frequency information which is for example read out from the frequency storage unit 630. A control unit 616 determines from the estimation values of the variation in time of the residual oscillation the synchronisation information based on the actual-phase and the actual-amplitude. In this process the control unit 616 takes the retardation resulting from the filtering, the required calculations and transient oscillation retardations into account. The control unit 616 calculates the initial phase φ_0 and the initial value A_s for the amplitude of the harmonic set point signal and the time t_0 at which the harmonic set point signal calculated in this way is in phase with the actual oscillation of the oscillator 190.

The synchronisation unit 620 comprises for example a controllable oscillation circuit 622, whose phase is controllable. According to an embodiment also the frequency of the oscillator is controllable. For example, the frequency of the oscillator is temperature-controlled such that a temperature-dependent change of the harmonic resonance angular frequency of the oscillator 190 may be followed. According to another embodiment the resonance angular frequency ωr of the oscillator circuit 622 is determined by the last entry into the frequency storage unit 630. The information about the last stored frequency may be supplied to the oscillator circuit 622 directly from the frequency storage unit 630 or via the control unit 616.

The amplitude of the harmonic set point signal is controlled such that it is ramped up according to a time function r(t) based on an initial amplitude value A_S corresponding to the estimated actual-amplitude of the oscillation of the oscillator 190 to the set value of the amplitude of the harmonic oscillation of the oscillator 190. For example the control unit 616 outputs in this process a ramp signal with the initial value of the estimated actual amplitude and the end value of the set point amplitude, whose variation in time and/or whose time constant is adapted to the actual phase.

A multiplicator unit 626 multiplies the amplitude signal $A_S \cdot r(t)$ with the output signal of the oscillator circuit 622. At the summation point 221 the difference between the harmonic set point signal and the measurement signal is formed. According to an embodiment the control unit 616 deactivates the filter unit 612 once their results are transmitted to the control unit 616. The deactivation of the filter unit 612 reduces for example the power consumption. In addition in a realization in a microprocessor the computation capacity needed for the estimation may be made free for the computation operations necessary during normal operation of the device. According to another embodiment, the control unit 616 controls the actuator activation unit 695 of Figure 1 such that the actuator unit 180 is switched on at time t_0.

According to an embodiment the controller extension unit 600 comprises a pre-stage unit 640. The pre-stage unit 640 determines from the measurement signal whether the amplitude A of the residual oscillation falls below a minimal threshold A_{xmin}. If this is the case it is to be assumed that the oscillator does not perform any significant residual movements anymore and that the oscillator may be started without problems from its resting state. According to an embodiment the harmonic set point signal starts then with the initial amplitude $A_S = A_{xmin}$. The phasing during start from the state of rest is arbitrary and the frequency of the harmonic set point signal may be determined for example from the knowledge of the temperature and a linear temperature model for the oscillator 190 or may be read out from the frequency storage unit 630. According to an embodiment the pre-stage unit 640 determines a maximal value A_{max} from several oscillation periods and a minimal value A_{min} and calculates from the values A_{max} and A_{min} rough values for the current oscillation amplitude A and a current oscillation offset A_0 according to equations 1 and 2:

(1) $$A = \frac{A_{max} - A_{min}}{2}$$

(2) $$A_0 = \frac{A_{max} + A_{min}}{2}$$

The controller extension unit 600 is for example realized as a digital circuit, for example as ASIC (application specific integrated circuit), DSP (digital signal processor) or FPGA (field programmable gate array). Controller extension unit 600 and controller main unit 200 may be formed in the same or in different components. According to a further embodiment the controller extension unit 600 and the controller main unit 200 are completely or partly programs, which are performed from a computer or a microprocessor.

According to an embodiment the filter unit 612 is a Kalman-filter. The estimation of values for amplitude and phase is performed in comparison to the decay time constant of the oscillator several magnitudes faster. According to an embodiment the movement y0 of the oscillator is therefore assumed to be an undamped harmonic oscillation:

(3) $$y_0(t) = A \times \sin(w_{0M} \times t)$$

The measurement signal y*(t) contains beside the actual oscillator movement y0 also the unavoidable measurement noise w and the a constant offset A0.

(4) $$y^* = y + A_0 = y_0 + w + A_0$$

The oscillator movement y0(t) is considered as the solution of the differential equation system with the equations (5) to (10):

(5) $\dot{y}_0 = A \cdot \omega_{0M} \cdot \cos(\omega_{0M} \cdot t)$

(6) $\ddot{y}_0 = -A \cdot \omega_{0M}{}^2 \cdot \sin(\omega_{0M} \cdot t)$

(7) $x_1 = y_0$

(8) $x_2 = \dot{y}_0$

(9) $\dot{x}_1 = x_2$

(10) $\dot{x}_2 = -\omega_{0M}{}^2 \cdot x_1$

The description of state of this system can be represented by using matrix notation as follows with equation (11):

(11) $$\underline{\dot{x}} = \begin{bmatrix} 0 & 1 \\ -\omega_{0M}{}^2 & 0 \end{bmatrix} \cdot \underline{x} = \underline{A} \cdot \underline{x}$$

In order to obtain a difference equation system, the above system is discretized with respect to the used sampling T of the measurement signal. This is represented by means of the Laplace transformation:

(12) $$\underline{\phi}(s) = \left[s \cdot \underline{I} - \underline{A} \right]^{-1}$$

(13) $$\underline{\phi}(s) = \begin{bmatrix} s & -1 \\ \omega_{0M}{}^2 & s \end{bmatrix}^{-1} = \begin{bmatrix} \dfrac{s}{s^2 + \omega_{0M}{}^2} & \dfrac{1}{s^2 + \omega_{0M}{}^2} \\ \dfrac{-\omega_{0M}{}^2}{s^2 + \omega_{0M}{}^2} & \dfrac{s}{s^2 + \omega_{0M}{}^2} \end{bmatrix}$$

(14) $$\underline{\phi}(t) = \begin{bmatrix} \cos(\omega_{0M} \cdot t) & \dfrac{1}{\omega_{0M}} \cdot \sin(\omega_{0M} \cdot t) \\ -\omega_{0M} \cdot \sin(\omega_{0M} \cdot t) & \cos(\omega_{0M} \cdot t) \end{bmatrix}$$

(15) $$\underline{f}(T) = \begin{bmatrix} \cos(\omega_{0M} \cdot T) & \frac{1}{\omega_{0M}} \cdot \sin(\omega_{0M} \cdot T) \\ -\omega_{0M} \cdot \sin(\omega_{0M} \cdot T) & \cos(\omega_{0M} \cdot T) \end{bmatrix}$$

Element for element the difference equation system may be represented as follows:

(16) $$x_1(k+1) = \cos(\omega_{0M} \cdot T) \cdot x_1(k) + \frac{1}{\omega_{0M}} \cdot \sin(\omega_{0M} \cdot T) \cdot x_2'(k)$$

(17) $$x_2'(k+1) = -\omega_{0M} \cdot \sin(\omega_{0M} \cdot T) \cdot x_1(k) + \cos(\omega_{0M} \cdot T) \cdot x_2'(k)$$

In order to simply the calculation and implementation, it is preferable to normalize the state variables x_2' with ω_{0M} as follows:

(18) $$x_1(k+1) = \cos(\omega_{0M} \cdot T) \cdot x_1(k) + \sin(\omega_{0M} \cdot T) \cdot \frac{x_2'(k)}{\omega_{0M}}$$

(19) $$\frac{x_2'(k+1)}{\omega_{0M}} = -\sin(\omega_{0M} \cdot T) \cdot x_1(k) + \cos(\omega_{0M} \cdot T) \cdot \frac{x_2'(k)}{\omega_{0M}}$$

With the new state variable

(20) $$x_2 = \frac{x_2'}{\omega_{0M}}$$

the description of state of the discretized system may be represented as follows:

(21) $$\underline{x}(k+1) = \underline{f}(T) \cdot \underline{x}(k) = \begin{bmatrix} \cos(\omega_{0M} \cdot T) & \sin(\omega_{0M} \cdot T) \\ -\sin(\omega_{0M} \cdot T) & \cos(\omega_{0M} \cdot T) \end{bmatrix} \cdot \underline{x}(k)$$

(22) $$y(k) = \underline{c}^T \cdot \underline{x}(k) + w(k) = \begin{bmatrix} 1 & 0 \end{bmatrix} \cdot \underline{x}(k) + w(k)$$

For the variance of the measurement noise w the symbol R is used in what follows.

Based on the noisy measurement values y* according to an embodiment a Kalman-filter is used in order to gain an estimation value $\hat{X}$ of the actual system state $\underline{x}$. The Kalman filter may be described by the following set of equations:

(23) $$\underline{P}^*(0) = \begin{bmatrix} 1/2 & 0 \\ 0 & 1/2 \end{bmatrix} \underline{x}^*(0) = \begin{bmatrix} 0 \\ 0 \end{bmatrix}$$

(24) $$\underline{k}(k) = \underline{P}^*(k) \times \underline{c} \times \left\{ \underline{c}^T \times \underline{P}^*(k) \times \underline{c} + R \right\}^{-1}$$

(25) $$\underline{\tilde{P}}(k) = \underline{P}^*(k) - \underline{k}(k) \times \underline{c}^T \times \underline{P}^*(k)$$

(26) $$\underline{P}^*(k+1) = \underline{f} \times \underline{\tilde{P}}(k) \times \underline{f}^T$$

(27) $$y(k) = y^*(k) - A_0$$

(28) $$\underline{\hat{x}}(k) = \underline{x}^*(k) + \underline{k}(k) \times \left\{ y(k) - \underline{c}^T \times \underline{x}^*(k) \right\}$$

(29) $$\underline{x}^*(k+1) = \underline{f} \times \underline{\hat{x}}(k)$$

By considering element by element the Kalman filter equations may be represented by equations (30) to (42) as follows. Here, the symmetry of the matrices $\underline{P}^*$ and $\tilde{P}$ has been used:

(30) $$k_1(k) = \frac{P_{11}^*(k)}{P_{11}^*(k) + R}$$

(31) $$k_2(k) = \frac{P_{12}^*(k)}{P_{11}^*(k) + R}$$

(32) $$\tilde{P}_{11}(k) = P_{11}^*(k) - P_{11}^*(k) \times k_1(k)$$

(33) $$\tilde{P}_{12}(k) = P_{12}^*(k) - P_{11}^*(k) \times k_2(k)$$

(34) $$\tilde{P}_{22}(k) = P_{22}^*(k) - P_{12}^*(k) \cdot k_2(k)$$

(35)

$$P_{11}^{*}(k+1)=\cos^{2}(w_{0M}\times T)\times\tilde{P}_{11}(k)+2\times\cos(w_{0M}\times T)\times\sin(w_{0M}\times T)\times\tilde{P}_{12}(k)+\sin^{2}(w_{0M}\times T)\times\tilde{P}_{22}(k)$$

(36)

$$\begin{aligned}P_{12}^{*}(k+1)=&-\cos(w_{0M}\times T)\times\sin(w_{0M}\times T)\times\tilde{P}_{11}(k)+(\cos^{2}(w_{0M}\times T)-\sin^{2}(w_{0M}\times T))\times\tilde{P}_{12}(k)\\&+\cos(w_{0M}\times T)\times\sin(w_{0M}\times T)\times\tilde{P}_{22}(k)\end{aligned}$$

(37)

$$P_{22}^{*}(k+1)=\sin^{2}(w_{0M}\times T)\times\tilde{P}_{11}(k)-2\times\cos(w_{0M}\times T)\times\sin(w_{0M}\times T)\times\tilde{P}_{12}(k)+\cos^{2}(w_{0M}\times T)\times\tilde{P}_{22}(k)$$

(38) $$y(k)=y^{*}(k)-A_{0}$$

(39) $$\hat{x}_{1}(k)=x_{1}^{*}(k)+k_{1}(k)\times(y(k)-x_{1}^{*}(k))$$

(40) $$\hat{x}_{2}(k)=x_{2}^{*}(k)+k_{2}(k)\times(y(k)-x_{1}^{*}(k))$$

(41) $$x_{1}^{*}(k+1)=\cos(w_{0M}\times T)\times\hat{x}_{1}(k)+\sin(w_{0M}\times T)\times\hat{x}_{2}(k)$$

(42) $$x_{2}^{*}(k+1)=-\sin(w_{0M}\times T)\times\hat{x}_{1}(k)+\cos(w_{0M}\times T)\times\hat{x}_{2}(k)$$

The zero crossing from negative to positive values of the estimated signals $\hat{x}_1$ is used to start the set signal of the amplitude control in proper phase. The time zero is chosen such that it coincides with the detected zero crossing. The estimated signal is then proportional to $\sin(\omega_{0M}\cdot T\cdot k)$.

The set point signal of the amplitude control should start exactly in the zero crossing of the estimated signal with the initial phase 0. But as only sampling values with a temporal resolution of T are available, the zero crossing may not be met exactly in general. According to an embodiment for this reason a phase $\Delta\varphi_0$ to be taken into account additionally is determined from the sampling value $\hat{x}_{1.0}$ previous to the zero crossing and the sampling value $\hat{x}_{1.1}$ after the zero crossing by linear interpolation according to equation (43):

(43) $$\Delta\varphi_{0}=\frac{\hat{x}_{1,1}}{\hat{x}_{1,1}-\hat{x}_{1,0}}\cdot\omega_{0M}\cdot T$$

This phase becomes effective only in the following cycle, which leads to a retardation to be considered additionally. Moreover, further retardations depending on realisation have to be considered during signal processing (e.g. during capturing of measurement values). These further retardations in multiples of the sampling time are assumed to be n_{osc}. The necessary initial phase φ_0 is then according to equation (44):

$$(44) \qquad \varphi_0 = \Delta\varphi_0 + (n_{osc} + 1) \cdot \omega_{0M} \cdot T = \frac{\hat{x}_{1,1}}{\hat{x}_{1,1} - \hat{x}_{1,0}} \cdot \omega_{0M} \cdot T + (n_{osc} + 1) \cdot \omega_{0M} \cdot T$$

Figure 2B illustrates details of a Kalman-filter as filter unit 612 of a control device.

If the system consists of several oscillators and the drives of the oscillators have to be started together because of a common electronic (e.g. in an inertial measurement unit (IMU) consisting of three rotation rate sensors) the zero crossings of the oscillators do not have to happen within the same sampling cycle necessarily. For this reason the starting phase for those oscillators, which have already had their zero crossings, is increased by $\omega_{oM} \cdot T$ for each additional sampling cycle. This is performed until also the last oscillator has had its zero crossing.

For example, as soon as the synchronisation condition (the zero crossing of the measured oscillator signal) has been detected for at least one of these sensors, the drive of this sensor is started according to the above-described method. The switch-on of the remaining sensors may then be performed as soon as also their individual synchronisation conditions are satisfied. The time for switching on the drive is chosen for all sensors individually in this process. According to another embodiment with a common electronic for which only a common switch-on time of the drive is possible, it is waited until the synchronisation conditions for all sensors are detected. As the synchronisation conditions per sampling cycle can be evaluated once, for each sensor, whose synchronisation condition was already obtained, the phase $\omega_{0M} \cdot T$ corresponding to the sampling cycle is added to the start phase for each additional sampling cycle.

The embodiment illustrated in Figure 3A describes in order to clarify the mode of operation of the principle on which the controller main unit 200 is based an analogue embodiment of the controller main unit 200 within a device 100. The controller main unit 200 comprises a PI controller 225 for harmonic command variables or harmonic set point signals with an integrating transfer element 222 with an integral action coefficient K_i and a proportional transfer element 224 with an amplification factor K_p. The PI controller 225 for harmonic command variables generates from a by the step function amplitude modulated harmonic oscillation of constant amplitude at the controller input a harmonic oscillation with the same frequency and time proportional amplitude at the controller output.

Figure 3B illustrates the transformation of a sine wave modulated step function signal $x_d(t)$ into a harmonic output signal u(t) with time proportional amplitude by the transfer function $G_{R0}(s)$ of the PI-controller 225. The described behavior of the PI-controller requires a dimensioning of the controller parameters K_i, K_p as described subsequently. Equation (45) gives the relation between the controller output signal u(t) and the controller input signal $x_d(t)$ for $x_d(t) = \sigma(t)$:

(45) $$u(t) = (K_P + K_I \cdot t) \cdot \sin(\omega_0 \cdot t) \cdot \sigma(t)\,.$$

The Laplace-transforms of the controller output signal u(t) and controller input signal $x_d(t)$ result from equations (46a) and (46b):

(46a) $$X_d(s) = \frac{\omega_0}{s^2 + \omega_0^2}$$

(46b) $$U(s) = K_P \cdot \frac{\omega_0}{s^2 + \omega_0^2} + K_I \cdot \frac{2 \cdot \omega_0 \cdot s}{(s^2 + \omega_0^2)^2}\,.$$

The transfer function $G_{R0}(s)$ of the PI-controller 225 for harmonic set point signals results thus from equation (47):

(47) $$G_{R0}(s) = \frac{U(s)}{X_d(s)} = K_P \cdot \frac{s^2 + 2 \cdot \frac{K_I}{K_P} \cdot s + \omega_0^2}{s^2 + \omega_0^2}$$

Characteristically for the continuous PI-controller 225 is a conjugate complex pole at $s = \pm j\omega_0$ resulting from the generalized integral component. With a harmonic oscillation of the frequency ω_0 at the controller input, the PI-controller 225 generates no phase shift at the controller output. For adjusting of an arbitrary phase the controller main unit 200 includes therefore additionally a dead time element 226 with the controller dead time T_R in series to the PI-controller 225. The controller transfer function $G_R(s)$ of the controller main unit 200 results thus from equation (48):

(48) $$G_R(s) = G_{R0}(s) \cdot e^{-T_R \cdot s} = K_P \frac{s^2 + 2 \cdot \frac{K_I}{K_P} \cdot s + \omega_0^2}{s^2 + \omega_0^2} \cdot e^{-T_R \cdot s}$$

The controller parameters K_i, K_p are chosen such that the controller zeros in the controller transfer function according to equation (48) compensate the conjugate complex system pole in the system transfer function according to equation (49).

(49) $$G(s) = \frac{A}{(s + s_o)^2 + \omega_0^2} \cdot e^{-T_S \cdot s} = G_o(s) \cdot e^{-T_S \cdot s}$$

By equating the coefficients of equations (48) and (49) for the determination of the controller parameters K_i, K_p the equations (50a) and (50b) result:

(50a) $$2 \cdot \frac{K_I}{K_P} \overset{!}{=} 2 \cdot s_0$$

(50b) $$\omega_0^2 \overset{!}{=} \omega_0^2 + s_0^2 .$$

According to one embodiment the damping s_0 and the resonance angular frequency ω_0 of the oscillator 190 are chosen such that $s_0 << \omega_0$ is satisfied and that hence equation (50b) is satisfied in very good approximation. From equation (50a) equation (50c) results as dimensioning rule for the ratio of the integral action coefficient K_I to the amplification factor K_P:

(50c) $$\frac{K_I}{K_P} \overset{!}{=} s_0 .$$

The transfer function $G_k(s)$ of the corrected open loop results from the product of the system transfer function $G_S(s)$ and the controller transfer function $G_R(s)$. As the expressions for the conjugate complex system pole and the conjugate complex controller zero cancel away by appropriate dimensioning according to equations

(50b), (50c), the transfer function $G_k(s)$ of the corrected open loop results from equation (51):

(51) $$G_k(s) = G_S(s) \cdot G_R(s) = A \cdot K_P \cdot \frac{1}{s^2 + \omega_0^2} \cdot e^{-(T_S + T_R) \cdot s}$$

By feedback control with a conventional PI-controller a phase jump from +90° to -90° occurs in the phase frequency response of the corrected open loop at the frequency $\omega = 0$. In contrast in the PI-controller 225 designed for harmonic command variables a 180° phase jump occurs at the frequency ω_0, which is however not necessarily between +90° and -90°. According to one embodiment the controller dead time T_R is therefore chosen such that the 180° phase jump occurs as much as possible exactly at ω_0, for example by dimensioning the controller parameters according to equation (52a):

(52a) $$(T_S + T_R) \cdot \omega_0 = \frac{3}{2} \cdot \pi$$

Is the phase shift produced by the system dead time T_S alone at ω_0 smaller than 90°, then the phase ratio of 180° can also be generated by an inverting controller. In this case the phases produced by the controller dead time T_R and the system dead time T_S at ω_0, respectively, have to add merely to $\pi/2$. The dimensioning rule for the controller dead time T_R is then:

(52b) $$(T_S + T_R) \cdot \omega_0 = \frac{\pi}{2}.$$

From the frequency response of the corrected open loop the stability properties of the closed loop can be deduced via the Nyquist criterion. The corrected open loop consists of the generalized integrator and the combination of system dead time T_S and controller dead time T_R. By appropriate dimensioning of the controller dead time T_R according to equations (52a) or (52b) the phase characteristics at the frequency ω_0 has a 180° jump between +90° for lower frequencies $\omega < \omega_0$ to -90° to higher frequencies $\omega > \omega_0$. The transfer function $G_w(s)$ of the closed loop results from the one of the corrected open loop $G_k(s)$.

(53) $$G_w(s) = \frac{G_k(s)}{1 + G_k(s)}.$$

When the controller dead time T_R is determined according to equation (52a) the closed loop is exactly then stable when the locus of the corrected open loop neither encloses nor runs through the point -1 for $0 \leq \omega < \omega_0$.

When in contrast the controller dead time T_R is determined according to equation (52b) and when the PI-controller 225 generates a 180° phase the closed loop is exactly then stable when the locus of the corrected open loop at a negative real axis starts at a value larger than -1.

In the interval $0 \leq \omega < \omega_0$ the absolute value characteristic intersects the 0 dB line at the gain crossover frequency, wherein the frequency distance to ω_0 at the gain crossover frequency determines the bandwidth of the closed loop. Via the amplification factor K_P the absolute value frequency response and hence the gain crossover frequency can be shifted along the ordinate such that the resulting bandwidth of the closed loop is adjustable. According to one embodiment the amplification factor K_P is chosen such that the bandwidth is maximal within the limits given by the stability criteria.

In summary, the controller main unit 200 comprises a PI-controller 225 for harmonic command variables, which is supplied with a harmonic set point signal and which comprises a proportional transfer element 224 with an amplification factor K_P and an integrating transfer element 222 with an integral action coefficient K_I. The integral action coefficient K_I and the amplification factor K_P are chosen such that in the s-plane the zero of the controller transfer function of the PI-controller 225 and the conjugate complex pole of the oscillator 190 describing the system transfer function compensate.

According to one embodiment the damping s_0 of the oscillator 190 with respect to the deflection in the direction of excitation is very much smaller than the resonance angular frequency ω_0 of the oscillator 190 and the ratio of the integral action coefficient K_I to the amplification factor K_P in sec^{-1} corresponds approximately to the damping s_0. Moreover, the amplification factor K_P can be chosen such that the resulting bandwidth is as high as possible for the respective stability requirements. The integral action coefficient K_I is then chosen in dependence from the damping s_o and the amplification factor K_P according to equation (50c).

According to one embodiment the system formed from the actuator unit 180, the oscillator 190 and the sensor unit 170 has a system dead time T_S and the controller main unit 200 has a dead time element 226 with the controller dead time T_R acting serially to the PI-controller 225. The controller dead time T_R is chosen in dependence of the resonance angular frequency ω_0 of the oscillator 190 and the system dead time T_S such that the phase frequency response of the corrected open loop at the frequency ω_0 has a phase jump from +90° to -90° towards higher frequencies.

According to a first variant of this embodiment the PI-controller for harmonic command variables does not flip the sign and the controller dead time T_R is chosen such that the product of the resonance angular frequency ω_0 and the sum of system dead time T_S and controller dead time T_R has $3\pi/2$ as a result. According to another variant of this embodiment the PI-controller 225 inverts the sign, respectively shifts the phase about 180°, and the phase effected by the controller dead time T_R and the system dead time T_S at the resonance angular frequency ω_0 merely adds to $\pi/2$ such that the product of the resonance angular frequency ω_0 and the sum of system dead time T_S and controller dead time T_R has $\pi/2$ as a result.

As the controller main unit 200 provides no baseband transformation, which requires a low pass filter for damping of higher frequency conversion products, the controller main unit 200 can be formed with a considerable broader band. The controller main unit 200 reacts faster to disturbances than comparative controllers which provide a baseband transformation.

Figures 4A and 4B refer to one embodiment at which the controller main unit 200 has a discrete PI-controller 325 for harmonic set point signals with a discrete proportional transfer element 324 with the amplification factor K_P and a discrete integrating transfer element 322 with the integral action coefficient K_I. According to an embodiment the analog measurement signal output by the sensor unit 170 is sampled with a sampling time T by a sampling unit 321 and converted into a digital input signal for the discrete PI-controller 325. According to another embodiment the sensor unit 170 outputs already a digital measurement signal.

According to an embodiment at which the system including the actuator unit 180, the oscillator 190 and the sensor unit 170 has a system dead time T_S, the controller main unit 200 includes a dead time element 326 arranged in series to the discrete PI-controller 325 with a controller dead time T_R. In what follows the system dead time T_S as well as the controller dead time T_R are expressed as multiplies of the sampling time T according to the equations (54a) and (54b):

(54a), (54b) $$T_S = \beta_S \cdot T \text{ and } T_R = \beta_D \cdot T .$$

In this process the controller dead time T_R is determined such that the corrected open loop has a phase jump at the resonance angular frequency ω_0 from +90° and -90° towards higher frequencies.

According to one embodiment the ratio of the integral action coefficient K_I to the amplification factor K_P is adjusted such that the controller zeroes of the controller transfer function compensate the conjugate complex system pole of the system transfer function in the s-plane. According to another embodiment the controller parameters are chosen such that the transfer function of the closed loop of an equivalent baseband system has a double real eigenvalue. The controller main unit 200 is for example realized as a digital circuit, for example as ASIC (application specific integrated circuit), DSP (digital signal processor) or FPGA (Field Programmable Gate Array) or as a program for a computer or micorprocessor.

Figure 4B illustrates the z-transfer function $G_{R0}(z)$ of the discrete PI-controller 325 for harmonic command variables according to Figure 4A. The transfer function $G_{R0}(z)$ is determined such that the PI-controller 325 generates from an input signal $x_d(k)$ including a harmonic oscillation modulated with the step function $\sigma(k)$ a harmonic oscillation of the same frequency with a time proportional amplitude as controller output signal u(k) as expressed by equation (55):

(55) $$u(k) = (K_P + K_I \cdot T \cdot k) \cdot \sin(\omega_0 \cdot T \cdot k) \cdot \sigma(k)$$

The input function $X_d(z)$ and the output function U(z) result from z-transformations according to equations (56a) and (56b):

(56a) $$X_d(z) = \frac{z \cdot \sin(\omega_0 \cdot T)}{z^2 - 2 \cdot \cos(\omega_0 \cdot T) \cdot z + 1}$$

(56b) $$U(z) = K_P \cdot \frac{z \cdot \sin(\omega_0 \cdot T)}{z^2 - 2 \cdot \cos(\omega_0 \cdot T) \cdot z + 1} + K_I \cdot \frac{T \cdot z^3 \cdot \sin(\omega_0 \cdot T) - T \cdot z \cdot \sin(\omega_o \cdot T)}{(z^2 - 2 \cdot \cos(\omega_0 \cdot T) \cdot z + 1)^2}$$

The transfer function $G_{R0}(z)$ of the discrete PI-controller 325 for harmonic command variables is then resulting from equation (56c):

(56c) $$G_{R0}(z) = \frac{U(z)}{X_d(z)} = \frac{(K_P + K_I \cdot T) \cdot z^2 - 2 \cdot K_P \cdot \cos(\omega_0 \cdot T) \cdot z + K_P - K_I \cdot T}{z^2 - 2 \cdot \cos(\omega_0 \cdot T) \cdot z + 1}$$

Because of the generalized integral portion such a discrete PI-controller has a pole at z = $e^{\pm j \cdot \omega_0 \cdot T}$ and generates with a harmonic oscillation of the frequency ω_0 at the input no phase shift at the output. To be able to nevertheless adjust an arbitrary phase, the controller main unit 200 is provided with a dead time element 326 with the retardation β_D according to one embodiment. The controller transfer function $G_R(z)$ of the controller main unit 200 with the dead time element 326 and the discrete PI-controller 325 result then from the equation (57):

(57) $$G_R(z) = G_{R0}(z) \cdot z^{-\beta_D} = \frac{(K_P + K_I \cdot T) \cdot z^2 - 2 \cdot K_P \cdot \cos(\omega_0 \cdot T) \cdot z + K_P - K_I \cdot T}{z^2 - 2 \cdot \cos(\omega_0 \cdot T) \cdot z + 1} \cdot z^{-\beta_D}.$$

The model of the continuous controlled system according to equation (49) has to be discretized accordingly. To this end in the transfer function G(s) of the controlled system according to equation (49) the system dead time T_S is at first expressed as a multiple of the sampling time T according to equation (54a):

(58) $$G(s) = \frac{A}{(s + s_0)^2 + \omega_0^2} \cdot e^{-\beta_S \cdot T_s} = G_0(s) \cdot e^{-\beta_S T_s}$$

Generally a step transfer function G(z) of a discretized model of a continuous controlled system with the transfer function G(s) can be calculated according to equation (59):

(59) $$G(z) = \frac{z - 1}{z} \cdot Z\left\{\frac{G(s)}{s}\right\}$$

With the following abbreviations according to equations (59a) to (59e)

(59a) $$K_S = \frac{A}{s_0^2 + \omega_0^2}$$

(59b) $$b_1 = 1 - e^{-s_0 \cdot T} \cdot \cos(\omega_0 \cdot T) - \frac{s_0}{\omega_0} \cdot e^{-s_0 \cdot T} \cdot \sin(\omega_0 \cdot T)$$

(59c) $$b_2 = e^{-2 \cdot s_0 \cdot T} - e^{-s_0 \cdot T} \cdot \cos(\omega_0 \cdot T) + \frac{s_0}{\omega_0} \cdot e^{-s_0 \cdot T} \cdot \sin(\omega_0 \cdot T)$$

(59d) $$a_1 = 2 \cdot e^{-s_0 \cdot T} \cdot \cos(\omega_0 \cdot T)$$

(59e) $$a_2 = -e^{-2 \cdot s_0 \cdot T}$$

the step transfer function G(z) for the oscillator 190 result from the equations (58) and (59) according to equation (60):

(60) $$G(z) = K_S \cdot \frac{b_1 \cdot z + b_2}{z^2 - a_1 \cdot z - a_2} \cdot \frac{1}{z^{\beta_s}} = G_0(z) \cdot \frac{1}{z^{\beta_s}}$$

According to one embodiment of the invention the controller dead time T_R is determined such that the phase frequency response of the corrected open loop has a phase jump from +90° to -90° towards higher frequencies at the resonance angular frequency ω_0. The z-transfer function for the corrected open loop results in analogy to equation (51) from the multiplication of the system transfer function G(z) according to equation (58) with the controller transfer function $G_R(z)$ according to equation (57):

(61) $$G_K(z) = G_0(z) \cdot G_{R0}(z) \cdot z^{-(\beta_S + \beta_D)}$$

Analog to the equations (52a) and (52b) the controller parameter β_D is chosen such that the transfer function of the corrected open loop $G_k(z)$ has a phase jump from +90° to -90° at the resonance angular frequency ω_0:

(62a) $$(\beta_S + \beta_D + \frac{1}{2}) \cdot \omega_0 \cdot T = \frac{3}{2} \cdot \pi$$

In comparison with equation (52a) one finds an additional part of $\frac{1}{2}\omega_0 T$ with respect to the continuous controller, which expresses a retardation, which can be traced back to the discretizing of an additional half sampling cycle. As in the case of the continuous controller a phase jump of 180° can be generated by a minus sign in the controller, provided that the phase shift generated by the system dead time $\beta_S \cdot T$ and the discretization, respectively, are smaller than 90° at the resonance angular frequency ω_0 such that the phases generated by the discretization, the controller dead time $\beta_D \cdot T$ and the system dead time $\beta_S \cdot T$, need merely to add up to $\pi/2$. Accordingly, the dimensioning rule for β_D results in this case from equation (62b):

(62b) $$(\beta_S + \beta_D + \frac{1}{2}) \cdot \omega_0 \cdot T = \frac{\pi}{2}.$$

The equations (62a) and (62b) lead normally to a non-integral value for β_D. Generally, the controller parameter β_D has an integral part n_D and a rest $1/a_D$ with $a_D > 1$ according to equation (63):

(63) $$\beta_D = n_D + \frac{1}{a_D}$$

According to one embodiment the integral part n_D can be approximated by a retardation chain in accordance with the length denoted by n_D and the fraction $1/a_D$ of a sampling cycle can be approximated by an all-pass filter of first order according to equation (64):

(64) $$z^{-\frac{1}{a_D}} \approx \frac{\alpha_D \cdot z + 1}{z + \alpha_D}$$

According to one embodiment the parameter α_D of the all-pass filter is chosen such that the phase of the exact transfer function $z^{-a_D^{-1}}$ and the phase of the all-pass approximation according to equation (64) coincide at the resonance angular frequency ω_0 as far as possible. From these conditions equation (65) results as a conditional equation for the parameter α_D of the all-pass filter:

(65) $$-\frac{\omega_0 \cdot T}{a_D} = \arctan\left(\frac{\alpha_D \cdot \sin(\omega_0 \cdot T)}{\alpha_D \cdot \cos(\omega_0 \cdot T) + 1}\right) - \arctan\left(\frac{\sin(\omega_0 \cdot T)}{\cos(\omega_0 \cdot T) + \alpha_D}\right)$$

According to one embodiment α_D is determined such that via nested intervals the zero of the function according to equation (66) is determined:

(66) $$f(\alpha_D) = \arctan\left(\frac{\alpha_D \cdot \sin(\omega_0 \cdot T)}{\alpha_D \cdot \cos(\omega_0 \cdot T) + 1}\right) - \arctan\left(\frac{\sin(\omega_0 \cdot T)}{\cos(\omega_0 \cdot T) + \alpha_D}\right) + \frac{\omega_0 \cdot T}{a_D}$$

The determination of n_D and a_D according to equations (63) and (66) is independent from the way of determining the further controller parameters K_P and K_I.

According to one embodiment of a method for manufacturing a controller unit which includes the dimensioning of the discrete PI-controller 325 according to Figure 4A, the amplification factor K_P and the integral action coefficient K_I of the discrete PI-controller 325 are chosen such that the controller zeros in the controller transfer function $G_R(z)$ according to equation (57) compensate the conjugate complex system pole of the system transfer function G(z) according to equation (60). Equating coefficients of equations (57) and (60) with respect to z^1 leads to the dimensioning rule according to equation (67):

(67) $$K_P \stackrel{!}{=} K_I \cdot T \cdot \frac{e^{-s_0 \cdot T}}{1 - e^{-s_0 \cdot T}}.$$

Equating coefficients with respect to z^0 leads to the dimensioning rule according to equation (68):

(68) $$K_P \stackrel{!}{=} K_I \cdot T \cdot \frac{1 + e^{-2 \cdot s_0 \cdot T}}{1 - e^{-2 \cdot s_0 \cdot T}}.$$

According to one embodiment the damping s_0 of the oscillator 190 and the sampling time T are chosen such that $s_0 \cdot T \ll 1$ holds such that the approximations according to (69a) and (69b) are sufficiently exact:

(69a) $$e^{-s_0 \cdot T} \approx 1 - s_0 \cdot T$$

(69b) $$e^{-2 \cdot s_0 \cdot T} \approx 1 - 2 \cdot s_0 \cdot T$$

With the approximations according to equations (69a) and (69b) the two independent dimensioning rules according to equations (67) and (68) can be approximated by a single dimensioning rule according to equation (70):

(70) $$K_P \stackrel{!}{=} K_I \cdot T \cdot \frac{1 - s_0 \cdot T}{s_0 \cdot T}, \text{ respectively } K_I \cdot T \stackrel{!}{=} K_P \cdot s_0 \cdot T.$$

According to one embodiment the ratio of the integral action coefficient K_I to the amplification factor K_P is set equal or nearly equal to the damping s_0 of the oscillator. The dimensioning of the discrete PI-controller 325 according to the described method which includes the compensation of the system pole by the controller zero, leads to a good reference action of the closed loop.

Figures 5A and 5B refer to one embodiment at which the controller main unit 200 has a controller extension 328, which is arranged in series to the PI-controller 325 and the dead time element 326 according to Figure 4A. In the following the structure of the controller extension 328 is deduced from an analogue controller extension for the baseband.

For example the oscillator 190 can have further resonances beside the resonance angular frequency at ω_0, such as mechanic structure resonances above or below the resonance angular frequency ω_0. The controller extension 328 is formed such that these further resonances are damped more strongly. To this end a retardation element of first order (PT_1-element) with a further pole at the kink frequency beyond the desired bandwidth would be added to a conventional PI-controller in the baseband. This additional controller pole effects that the controller is not any longer acting as a proportional element for high frequencies, but that its absolute value frequency drops down with 20 db/decade. The step response y(k) of such an extension in the baseband results from the step function σ(k) as input signal u(k) according to equation (71):

(71) $$y(k) = (1 - e^{-\frac{k \cdot T}{T_1}}) \cdot \sigma(k)$$

The z transform U(z) of the input signal u(k) corresponds to the z transform of the step signal:

(72a) $$U(z) = \frac{z}{z-1}$$

The z transform Y(z) of the output signal y(k) results from equation (72b):

(72b) $$Y(z) = \frac{z}{z-1} - \frac{z}{z - e^{-\frac{T}{T_1}}}$$

Analog to equation (47) for the transfer function $G_{RE0}(z)$ of such a controller extension in the baseband results thus:

(73) $$G_{RE0}(z) = \frac{1 - e^{-\frac{T}{T_1}}}{z - e^{-\frac{T}{T_1}}}$$

According to one embodiment the controller extension 328 in the bandpass band is configured now in analogy to the controller extension in the baseband such that the controller extension 328 responses to an admission with a harmonic oscillation of the resonance angular frequency ω_0 modulated by the step function with a harmonic oscillation of the same frequency, wherein the step response of the baseband extension defines the envelope as it is illustrated on the right side of Figure 5B.

Figure 5B illustrates the transformation of a sign modulated step function u(k) onto an output signal with a sign oscillation whose envelope results from the step response according to the transfer function $G_{RE0}(z)$ of the discrete controller extension in the bandpass band. The input signal of the controller extension 328 in the bandpass band with the transfer function $G_{RE}(z)$ results from equation (74):

(74) $$u(k) = \sin(\omega_0 \cdot T \cdot k) \cdot \sigma(k)$$

The controller output signal y(k) is a harmonic oscillation whose envelope corresponds to the step response of the PT_1-controller extension in the baseband:

(75) $$y(k) = (1 - e^{-\frac{k \cdot T}{T_1}}) \cdot \sin(\omega_0 \cdot T \cdot k) \cdot \sigma(k)$$

The z-transforms U(z) and Y(z) result from equations (76a) and (76b):

(76a) $$U(z) = \frac{z \cdot \sin(\omega_0 \cdot T)}{z^2 - 2 \cdot \cos(\omega_0 \cdot T) \cdot z + 1}$$

(76b) $$Y(z) = z \cdot \frac{\sin(\omega_0 \cdot T)}{z^2 - 2 \cdot \cos(\omega_0 \cdot T) \cdot z + 1} - z \cdot \frac{e^{-\frac{T}{T_1}} \cdot \sin(\omega_0 \cdot T)}{z^2 - 2 \cdot e^{-\frac{T}{T_1}} \cdot \cos(\omega_0 \cdot T) \cdot z + e^{-2\frac{T}{T_1}}}$$

The transfer function $G_{RE}(z)$ of the controller extension 328 for the bandpass band result from equation (77):

(77) $$G_{RE}(z) = \frac{Y(z)}{U(z)} = \frac{(1 - e^{-\frac{T}{T_1}}) \cdot z^2 - e^{-\frac{T}{T_1}} \cdot (1 - e^{-\frac{T}{T_1}})}{z^2 - 2 \cdot e^{-\frac{T}{T_1}} \cdot \cos(\omega_0 \cdot T) \cdot z + e^{-2\frac{T}{T_1}}}$$

The controller extension 328 with the transfer function $G_{RE}(z)$ acts in series to the discrete PI-controller 325 similarly to a bandpass of first order with the resonance angular frequency ω_0 as midband frequency. Absolute value and phase of the corrected open loop at the resonance angular frequency ω_0 in a narrow region around the resonance angular frequency ω_0 according to equation (78) remain unchanged.

(78) $$\omega_0 - \frac{1}{T_1} \leq \omega \leq \omega_0 + \frac{1}{T_1}$$

In this region the absolute value frequency response of the corrected open loop is hardly influenced, while out of this region a considerable drop of the absolute value happens such that possible undesired resonances can be dropped.

Figures 6A to 6C refer to a micromechanical rotation rate sensor 500 according to a further embodiment. The rotation rate sensor 500 includes an excitation unit 590, e.g. an excitation frame, suspended at first spring elements 541. The first spring elements 541 couple the excitation unit 590 to an attachment structure 551 which is fixedly connected to a support substrate 550 illustrated in Figure 6B. The spring elements 541 damp a deflection of the excitation unit 590 with respect to the support substrate 550 along the direction of excitation 501 only weakly. Via second spring elements 542 a detection unit 580 is coupled to the excitation unit 590 and is movable with respect to the excitation unit 590 mainly along a detection direction 502 orthogonal to the direction of excitation 501. The direction of excitation 501 and the detection direction 502 run parallel to a surface of the support substrate 550. The first and second spring elements 541, 542 are for example beam-like structures with small cross sections, which are formed between each of the structures to be coupled.

According to one embodiment the rotation rate sensor 500 includes first force transmission and sensor units 561, 571, e.g. electrostatic force transmitters and sensors, which excite the system formed from the excitation unit 590 and the detection unit 580 to an oscillation along the direction of excitation 501 and/or are able to capture a corresponding deflection of the excitation unit 590. The rotation rate sensor 500 includes further second force transmission and sensor units 562, 572, e.g. electrostatic force transmitters and sensors, which act on the detection unit 580 and/or are able to capture its deflection. According to one embodiment at least one of the second force transmission and sensor units 562, 572 is driven such that it counteracts a deflection of the detection unit 580, caused by a disturbance or in case of a closed loop system caused by a measured variable.

During operation of the rotation rate sensor 500 the first force transmission and sensor units 561, 571 excite for example the excitation unit 590 to an oscillation along the direction of excitation 501, wherein the detection unit 580 moves approximately with the same amplitude and phase with the excitation unit 590. When the arrangement is rotated around an axis orthogonal to the substrate plane a Coriolis force is acting on the excitation unit 590 and the detection unit 580, which deflects the detection unit 580 with respect to the excitation unit 590 in the detection direction 502. The second force transmission and sensor units 562, 572 capture the deflection of the deflection unit 580 and thus the rotational movement around the axis orthogonal to the substrate plane.

According to one embodiment at least one of the force transmission and sensor units 561, 572, 562, 572 acts as actuator and either the excitation unit 590 or the detection unit 580 as oscillator within the meaning of one of the devices 200 described above.

According to one embodiment illustrated in Figure 6C of the rotation rate sensor 500 for example the first force transmission and sensor units 561, 571 excite the excitation unit 590 to an oscillation with the resonance angular frequency ω_0 along the direction of excitation 501. In a control loop with a controller main unit 200 and a controller extension unit 600 according to the above discussion an oscillation of the detection unit 580 along the detection direction 502 (x2-oscillator) can then for example correspond to the harmonic force signal as described above.

The deflection of the x2-oscillator can be captured via the charge on the common movable electrode, which is formed on the excitation unit 590. The charge can be measured via one of the attachment structures 551. A charge amplification unit 521 amplifies the measured signal. While typically a demodulation unit demodulates the measured signal with a frequency which corresponds for example to the resonance angular frequency ω_0 before it is fed into a controller unit, the embodiments of the invention provide to feed the non-demodulated harmonic signal as measurement signal within the meaning described above into a controller unit 520 according to the above discussion.

The damping s_0 effective for the oscillation is considerably smaller than the resonance angular frequency ω_0. The signal measured via the excitation frame respectively the excitation unit 590 reproduces partly the movement of the excitation unit 590 along the direction of excitation 501. A disturbance whose source can be outside of the rotation rate sensor 500, or, in a closed loop system, the measurement variable, respectively, superposes the oscillation and modulates its amplitude. The controller unit 520 deduces from the modulated harmonic signal a control signal for the second force transmission and sensor units 562, 572 which effects that these counteract the deflection effected by the disturbance or the measurement variable, respectively. An amplification unit 522 transforms the control signal in a suitable reset signal for the electrodes of the second force transmission and sensor units 562, 572. The controller unit 520 comprises one of of the controller main units 200 and controller extension units 600, respecitvely, described above.

The rotation rate sensor 505 illustrated in Figure 7 differs from the rotation rate sensor 500 illustrated in Figure 6A by a Coriolis unit 585 arranged between the excitation unit 590 and the detection unit 580. Second spring elements 542 which couple the Coriolis unit 585 to the excitation unit 590 allow for a deflection of the Coriolis unit 585 relative to the excitation unit 590 in the detection direction 502. Third spring elements 543, which can be connected partly with the support substrate 550, couple the detection unit 580 to the Coriolis unit 585 such that the detection unit 580 can follow the movement of the Coriolis unit 585 along the detection direction 502, but cannot follow movements along the direction of excitation 501. The detection unit 580 is fixed with respect to the direction of excitation 501 and is moveable along the detection direction 502.

According to other embodiments at least one of the first or second force transmission and sensor units 561, 562, 571, 572 acts as actuator and either the excitation unit 590 or the detection unit 580 or the excitation unit 590 as well as the detection unit 580 act as oscillator according to one of the devices described above, which are operated according to the principle of the bandpass controller. In this process the force transmission and sensor units 561 and 571 act as force transmission and sensor units respectively for the x1-oscillator and the force transmission and sensor units 562 and 572 act as force transmission and sensor units respectively for the x2-oscillator.

A rotation rate sensor according to another embodiment includes two of the arrangements as illustrated in Figure 6A or Figure 7, which are coupled to each other such that the excitation units perform opposing oscillations in the stationary state with respect to each other. Other embodiments concern rotation rate sensors with four of the arrangements as illustrated in Figure 6A or Figure 7 which are coupled to each other such that every two of the excitation units perform opposing oscillations in the stationary state.

A further embodiment refers to a rotation rate sensor device having the combination of controller main unit 200 and controller extension unit 600 as illustrated in Figures 1, 2A and 2B. The controller main unit 200 includes at least one PI-controller 225, 325 for harmonic set point signals, which on his part has a proportional transfer element 224, 324 and an integrating transfer element 222, 322 arranged in parallel to the proportional transfer element 224, 324, wherein a controller input of the controller main unit 200 is connected with both transfer elements 222, 224, 322, 324. The transfer function of the PI-controller 225, 325 for harmonic set point signals has a conjugate complex pole at a controller angular frequency ω_r in the s-plane or at $e^{\pm j\omega_r T}$ in the z-plane, wherein T is the sampling time of a discrete input signal of the PI-controller 325 and wherein ω_r is larger than 0.

To this end an integral action coefficient of the integrating transfer element 222, 322 and an amplification factor of the proportional transfer element 224, 324 is chosen such that the PI-controller 225, 325 for harmonic command variables is suitable for generating at a controller output a harmonic oscillation of the controller angular frequency ω_r with rising amplitude, at admission with an harmonic input signal of the controller angular frequency ω_r modulated by the step function at the controller input.

The PI-controller 225, 325 for harmonic set point signals can also be taken for a controller derived from a conventional PI-controller for stationary set point signals and differs from it by the position of the poles in the s- or z-plane, respectively.

Figure 8 refers to a method for operating a control device with harmonic command variable. A sensor unit generates a measurement signal, which represents the deviation of an oscillator along a direction of excitation (802). A controller extension unit generates, based on an estimation of an actual phase and actual amplitude of a residual oscillation of the oscillator at deactivated actuator unit (804), a phase synchronous harmonic set point signal with equal amplitude. A controller main unit generates from the measurement signal and the synchronous set point signal a synchronized control signal for the actuator unit such that the actuator unit can counteract against a deviation of the oscillator from a harmonic oscillation. At or after providing the synchronized control signal the actuator unit is activated (806).

Patentkrav

1. Reguleringsinnretning, omfattende
en sensorenhet (170) som er utformet for å gi ut et målesignal som avbilder et sideutslag av en oscillator (190) langs en eksitasjonsretning,
en aktuatorenhet (180) som innvirker på oscillatoren (190) og kan styres ved styringssignaler, **karakterisert ved**
en regulator-hovedenhet (200) som er utformet for å derivere et styringssignal for aktuatorenheten (180) fra målesignalet og fra et harmonisk skal-verdi-signal på en slik måte at aktuatorenheten (180) virker imot et avvik av oscillatorens (190) sideutslag fra en skal-amplitude til en harmonisk resonanssvingning av oscillatoren (190), og
en regulator-ekspansjonsenhet (600) som er utformet for å bestemme en faktisk fase og en faktisk amplitude til en restsvingning av oscillatoren (190) ut fra målesignalet når aktuatorenheten (180) er deaktivert, og for å utgi et harmonisk skal-verdi-signal som er tilpasset den faktiske fasen og den faktiske amplituden, til regulator-hovedenheten (200), slik at etter aktivering av aktuatorenheten (180) blir en amplitude til oscillatorens (190) restsvingning fasesynkront forsterket inntil en skal-amplitude, mens energi som er inneholdt i restsvingningen, blir utnyttet.

2. Reguleringsinnretning ifølge krav 1, **karakterisert ved**
at regulator-hovedenheten (200) er en regulatorinnretning (225, 325) for harmoniske skal-verdi-signaler, der reguleringsinnretningen (225, 325) omfatter minst ett proporsjonalt overføringsledd (224, 324) og ett integrerende overføringsledd (222, 322) anordnet parallelt med det proporsjonale overføringsleddet (224, 324), og en regulatorinngang til regulator-hovedenheten (200) er forbundet med begge overføringsleddene (222, 224, 322, 324), og
en overføringsfunksjon av regulatorinnretningen (225, 325) har en konjugert kompleks pol ved en regulatorkretsfrekvens ω_r i s-planet eller en pol ved $e^{\pm j\omega_r T}$ i z-planet, der T er avsøkingstiden til et diskret inngangssignal i regulatorinnretningen (225, 325) og ω_r er større enn 0.

3. Reguleringsinnretning ifølge krav 2, **karakterisert ved**
at en integreringskoeffisient til det integrerende overføringsleddet (222, 322) og en forsterkningsfaktor til det proporsjonale overføringsleddet (224, 324) er valgt slik at regulatorinnretningen (225, 325), når det påføres et

sprangfunksjonsmodulert harmonisk inngangssignal av regulatorkretsfrekvensen ω_r på regulatorinngangen, er utformet for å generere en harmonisk svingning av regulatorfrekvensen ω_r med økende amplitude på en regulatorutgang.

4. Reguleringsinnretning ifølge krav 3, **karakterisert ved at** integreringskoeffisienten og forsterkningsfaktoren er valgt slik at nullplassene til regulatorinnretningens (225, 325) overføringsfunksjon kompenserer poler til en overføringsfunksjon av oscillatoren (190).

5. Reguleringsinnretning ifølge et av kravene 1 til 4, **karakterisert ved at** regulator-ekspansjonsenheten (600) har:
en oppfangingsenhet (610) som er utformet for å bestemme den faktiske fasen og den faktiske amplituden til oscillatorens (190) restsvingning ut fra målesignalet, og bestemme en synkronisasjonsinformasjon for det harmoniske skal-verdi-signalet ut fra den faktiske fasen og den faktiske amplituden på en slik måte at det ved hjelp av aktuatorenheten (180) forsterkes en amplitude av oscillatorens (190) restsvingning fasesynkront, og energi som er inneholdt i restsvingningen, blir utnyttbar; og
en synkroniseringsenhet (620) som er utformet for å motta synkronisasjonsinformasjon og fastsette en fase og en start-amplitudeverdi for det harmoniske skal-verdi-signalet på basis av synkronisasjonsinformasjonen.

6. Reguleringsinnretning ifølge krav 5, **karakterisert ved** oppfangingsinnretningen (610) har:
en filterenhet (612) som er utformet for å fastslå estimerte verdier for et tidsforløp av restsvingningen ut fra målesignalet, fra en estimert verdi for den aktuelle svingningsfrekvensen, fra en estimert verdi for variansen til målestøy inneholdt i målesignalet og fra en estimert verdi for et konstant amplitude-offset av målesignalet; og
en styringsenhet (616) som er utformet for å fastslå, ut fra de estimerte verdiene for tidsforløpet av restsvingningen, synkronisasjonsinformasjonen som beskriver den faktiske fasen og den faktiske amplituden, og utgi den.

7. Reguleringsinnretning ifølge krav 6, **karakterisert ved at** styringsenheten (616) er utformet for å deaktivere filterenheten (612) etter utgivelse av synkronisasjonsinformasjonen.

8. Reguleringsinnretning ifølge et av kravene 6 eller 7, **karakterisert ved at** styringsenheten (616) er utformet for å aktivere aktuatorenheten (180) etter utgivelse av synkronisasjonsinformasjonen.

9. Reguleringsinnretning ifølge et av kravene 6 til 8, **karakterisert ved at** regulator-ekstensjonsenheten (600) er utformet for å øke amplituden til skal-verdi-svingningen innenfor et forhåndsdefinert tidsrom fra start-amplitudeverdien til en skal-amplitudeverdi.

10. Reguleringsinnretning ifølge et av kravene 1 til 9, **karakterisert ved at** filterenheten (612) er et Kalman-filter.

11. Reguleringsinnretning ifølge et av kravene 1 til 10, **karakterisert ved at** regulator-ekstensjonsenheten (600) omfatter en frekvenslagringsenhet (630) som er utformet for å lagre en frekvensinformasjon som beskriver en aktuell svingefrekvens av oscillatoren (190), i tidsavstander, og
regulator-ekstensjonsenheten (600) videre er utformet for å bruke den lagrede frekvensinformasjonen for estimering av den faktiske fasen og den faktiske amplituden til oscillatorens (190) restsvingning og/eller for generering av det harmoniske skal-verdi-signalet.

12. Reguleringsinnretning ifølge et av kravene 1 til 11, **karakterisert ved at** regulatorinnretningen er en del av en rotasjonsratesensor (500, 505), oscillatoren er utformet som en eksitasjonsenhet (590), en Coriolis-enhet (585) eller en deteksjonsenhet (580), og aktuatorenheten er en kraftgiver (561),
Coriolis-enheten (585) er festet på eksitasjonsenheten (590) slik at Coriolis-enheten (585) følger en bevegelse av eksitasjonsenheten (590) langs eksitasjonsretningen og i tillegg er bevegelig langs en deteksjonsretning som er vinkelrett på eksitasjonsretningen, og
deteksjonsenheten (580) er festet på eksitasjonsenheten (590) eller på Coriolis-enheten (585) slik at deteksjonsenheten (580) enten
følger en bevegelse av eksitasjonsenheten (590) langs eksitasjonsretningen og i tillegg er bevegelig langs en deteksjonsretning som er vinkelrett på eksitasjonsretningen, eller
følger en bevegelse av Coriolis-enheten (585) langs en deteksjonsretning som er vinkelrett på eksitasjonsretningen, og er fiksert langs eksitasjonsretningen.

13. Rotasjonsratesensor, omfattende
en bevegelig opplagret oscillator (190) som kan eksiteres til en harmonisk resonanssvingning i en eksitasjonsretning,
en sensorenhet (170) som er utformet for å gi ut et målesignal som avbilder et sideutslag av oscillatoren (190) langs eksitasjonsretningen,
en aktuatorenhet (180) som innvirker på oscillatoren (190) og kan styres ved styringssignaler, **karakterisert ved**
en regulator-hovedenhet (200) som er utformet for å derivere et styringssignal for aktuatorenheten (180) fra målesignalet og fra et harmonisk skal-verdi-signal på en slik måte at aktuatorenheten (180) virker imot et avvik av oscillatorens (190) sideutslag fra en skal-amplitude til resonanssvingningen, og
en regulator-ekspansjonsenhet (600) som er utformet for å bestemme en faktisk fase og en faktisk amplitude til en restsvingning av oscillatoren (190) ut fra målesignalet når aktuatorenheten (180) er deaktivert, og for å utgi et harmonisk skal-verdi-signal som er tilpasset den faktiske fasen og den faktiske amplituden, til regulator-hovedenheten (200), slik at etter aktivering av aktuatorenheten (180) blir en amplitude av oscillatorens (190) restsvingning fasesynkront forsterket inntil en skal-amplitude, mens energi som er inneholdt i restsvingningen, blir utnyttet.

14. Rotasjonsratesensor ifølge krav 13, der
oscillatoren er en eksitasjonsenhet (590) som ved hjelp av en kraftgiver (561) kan gis et sideutslag langs en eksitasjonsretning og er utformet for svingning med en resonanskretsfrekvens ω_0.

15. Fremgangsmåte for å drive en reguleringsinnretning med en harmonisk føringsstørrelse, omfattende
generering av et målesignal som avbilder et sideutslag til en oscillator (190) langs en eksitasjonsretning, ved hjelp av en sensorenhet (170),
generering av et fase- og amplitudesynkront harmonisk skal-verdi-signal basert på en estimering av en faktisk fase og en faktisk amplitude til en restsvingning av oscillatoren (190) ved deaktivert aktuatorenhet (180),
generering av et synkronisert styringssignal for en aktuatorenhet (180) ut fra målesignalet og det fase- og amplitudesynkrone harmoniske skal-verdi-signalet, slik at aktuatorenheten (180) virker imot et avvik av oscillatorens (190) sideutslag fra en harmonisk svingning, og

aktivering av aktuatorenheten (180) samtidig med eller etter tilveiebringelse av det synkroniserte styringssignalet.

16. Rotasjonsratesensor ifølge krav 13, **karakterisert ved at** regulator-hovedenheten (200) er en regulatorinnretning (225, 325) for harmoniske skal-verdi-signaler, der reguleringsinnretningen (225, 325) omfatter minst ett proporsjonalt overføringsledd (224, 324) og ett integrerende overføringsledd (222, 322) anordnet parallelt med det proporsjonale overføringsleddet (224, 324), og en regulatorinngang til regulator-hovedenheten (200) er forbundet med begge overføringsleddene (222, 224, 322, 324), og en overføringsfunksjon av regulatorinnretningen (225, 325) har en konjugert kompleks pol ved en regulatorkretsfrekvens ω_r i s-planet eller en pol ved $e^{\pm j\omega_r T}$ i z-planet, der T er avsøkingstiden til et diskret inngangssignal i regulatorinnretningen (225, 325) og ω_r er større enn 0.

Fig. 1

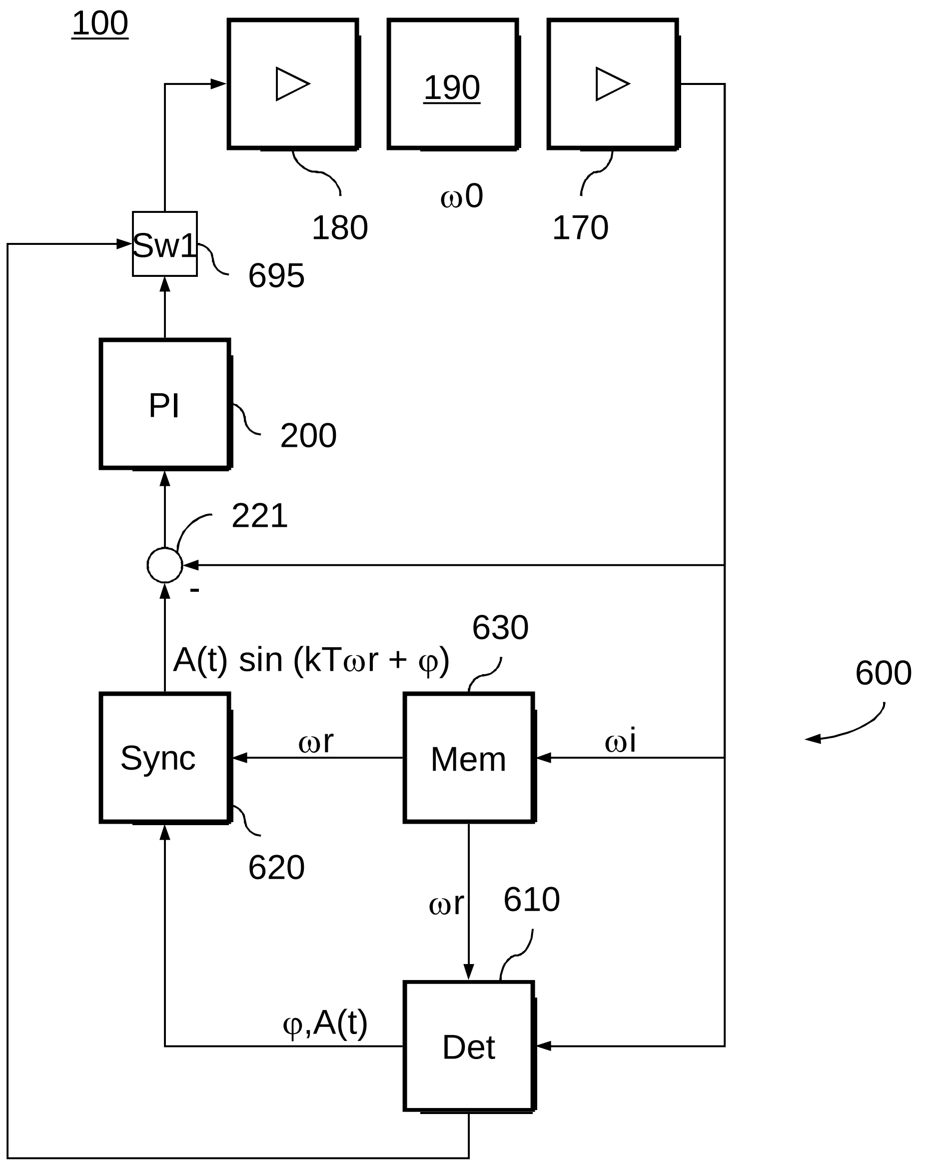

100
190
180
ω0
170
Sw1
695
PI
200
221
-
630
A(t) sin (kTωr + φ)
600
Sync
ωr
Mem
ωi
620
ωr
610
φ,A(t)
Det

Fig. 2A

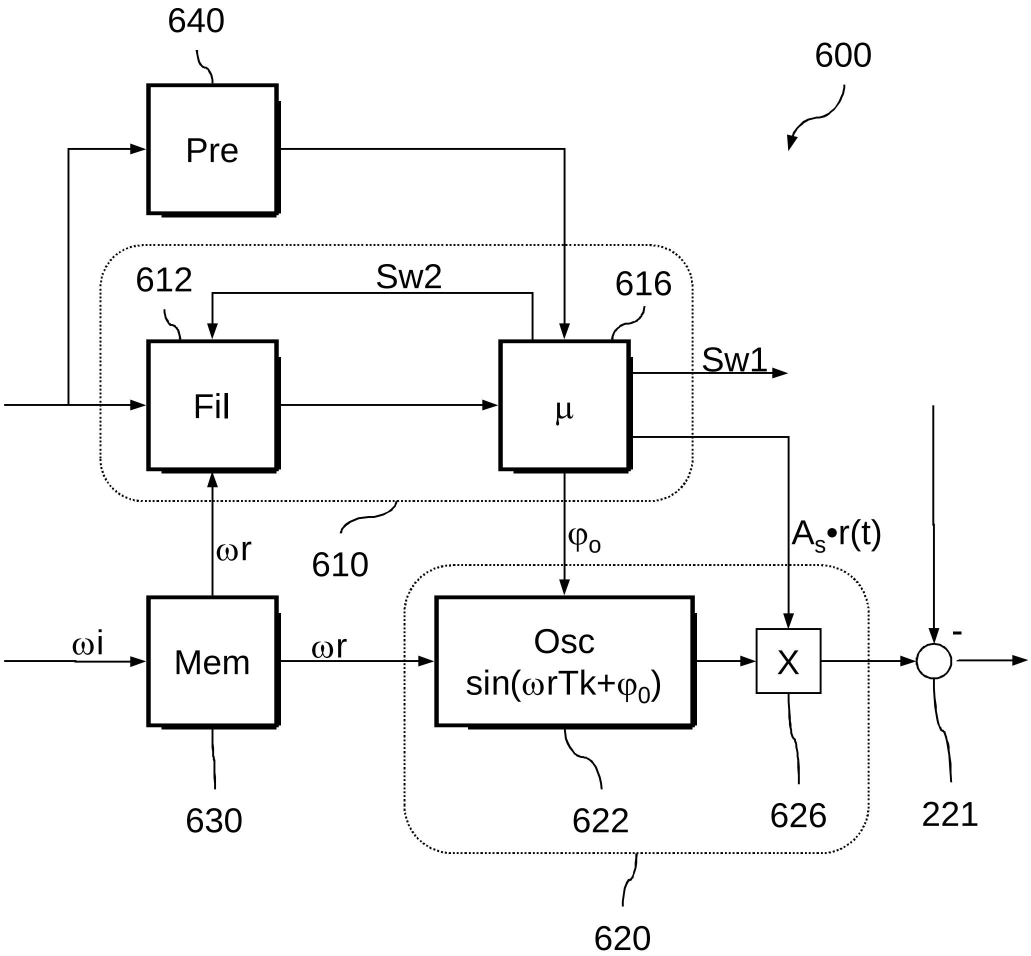

640
600
Pre
612
Sw2
616
Fil
Sw1
μ
ωr
610
φo
As•r(t)
ωi
Mem
ωr
Osc
sin(ωrTk+φ0)
X
-
630
622
626
221
620

Fig. 2B

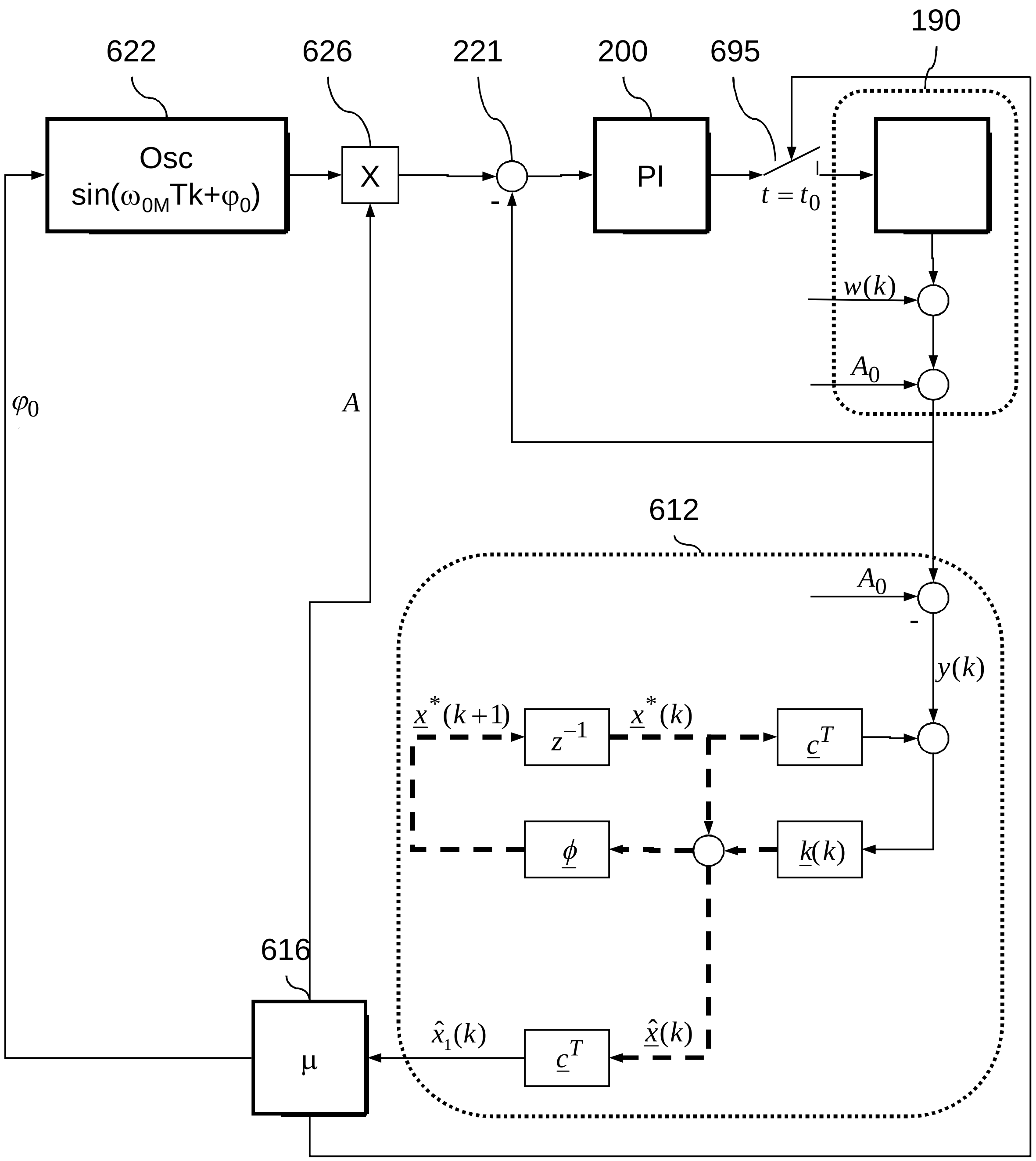
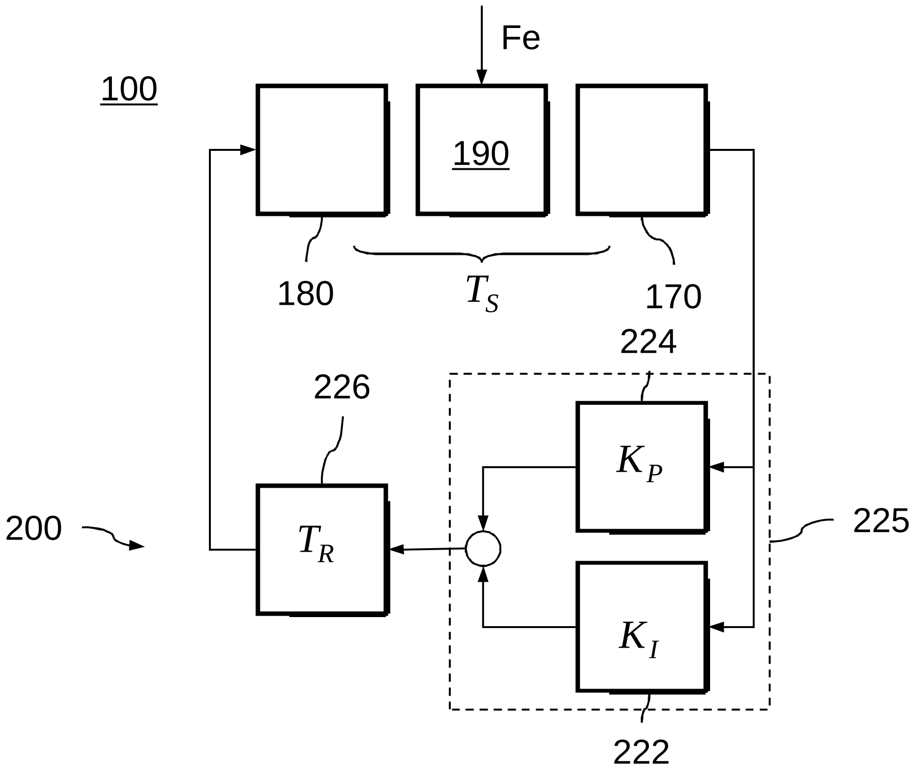

622
626
221
200
695
190
Osc
sin(ω0MTk+φ0)
X
PI
t = t0
w(k)
A0
φ0
A
612
A0
y(k)
x*(k+1)
z−1
x*(k)
cT
φ
k(k)
616
μ
x̂1(k)
cT
x̂(k)

Fig. 3A



$$\frac{K_I}{K_P} \approx s_0 \qquad vz = sign(K_I)$$

$$(T_S + T_R) \cdot \omega_0 = \frac{3}{2}\pi \quad \text{for} \quad vz = +1$$

$$(T_S + T_R) \cdot \omega_0 = \frac{1}{2}\pi \quad \text{for} \quad vz = -1$$

Fig. 3B

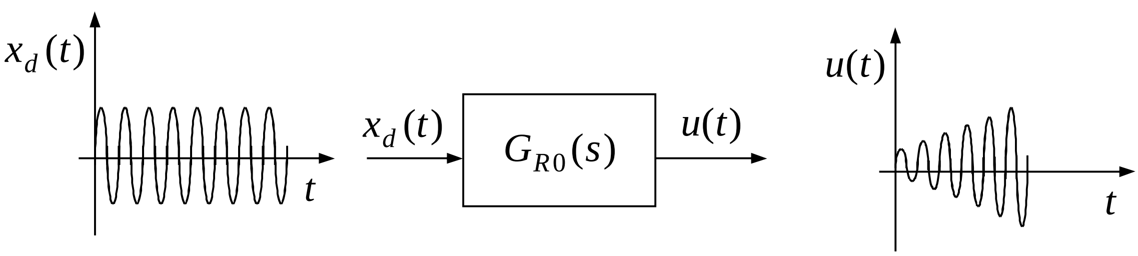
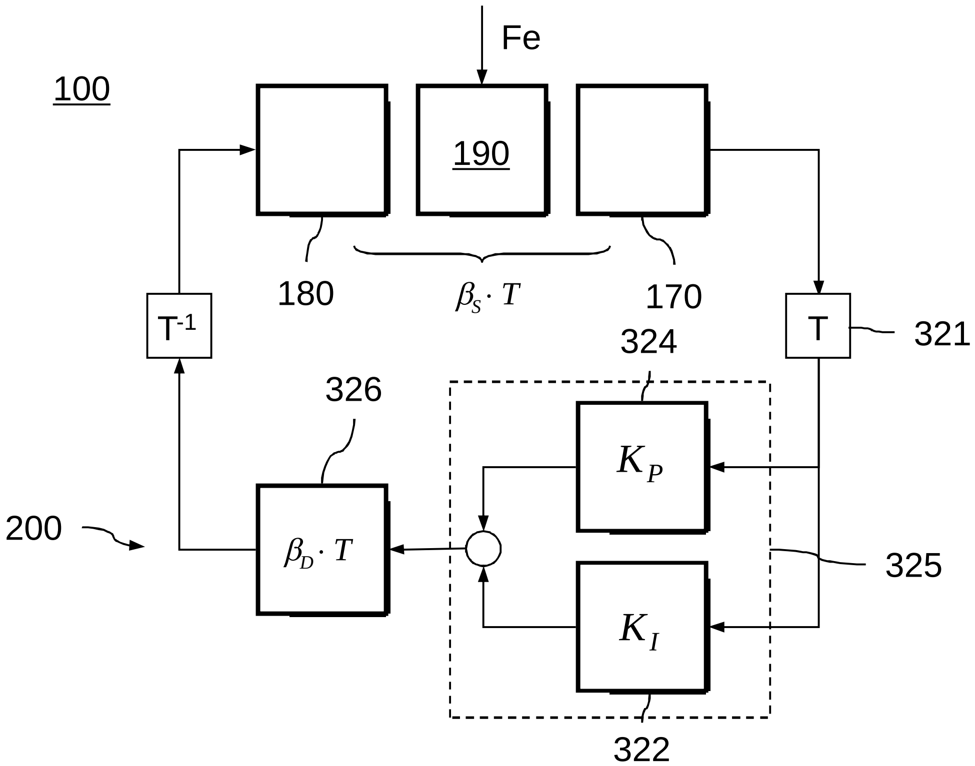


Fig. 4A



$$\frac{K_I}{K_P} \approx s_0 \quad vz = sign(K_I)$$

$$\left(\beta_S + \beta_D + \frac{1}{2}\right) \cdot \omega_0 \cdot T = \frac{3}{2}\pi \quad \text{for} \quad vz = +1$$

$$\left(\beta_S + \beta_D + \frac{1}{2}\right) \cdot \omega_0 \cdot T = \frac{1}{2}\pi \quad \text{for} \quad vz = -1$$

Fig. 4B

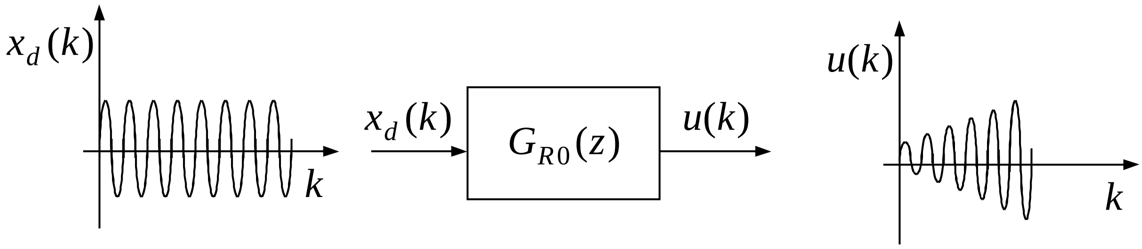


Fig. 5A

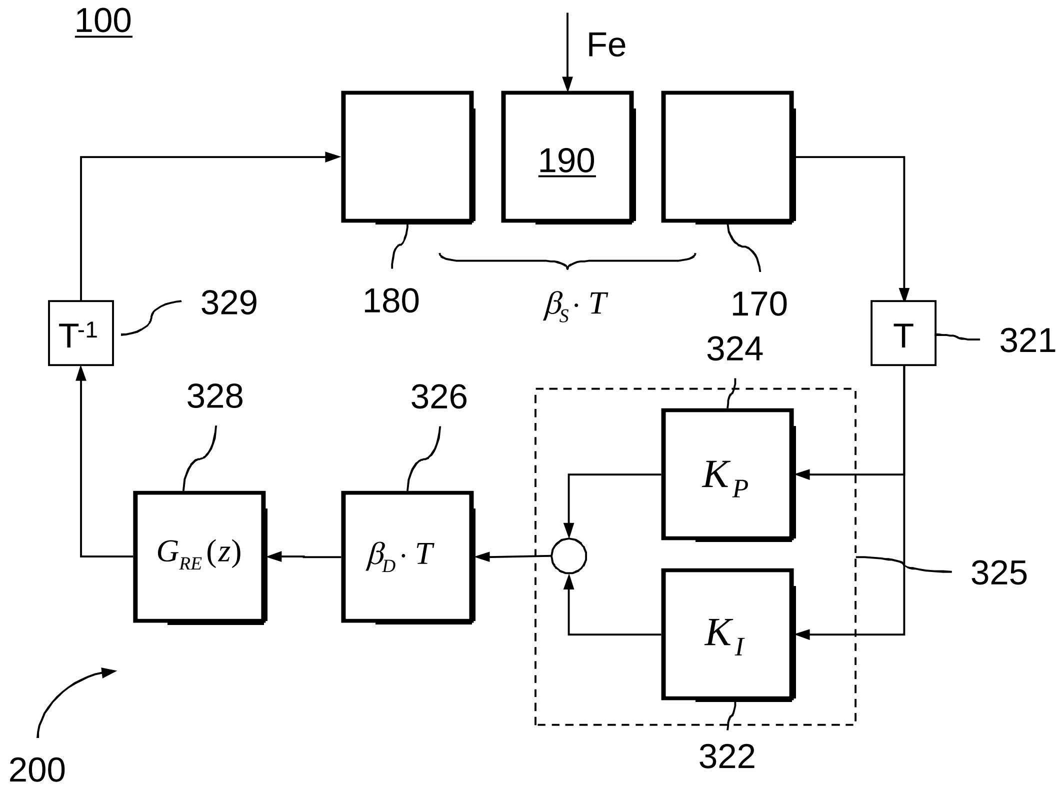

100
Fe
190
180
170
$\beta_S \cdot T$
329
T^{-1}
T
321
324
328
326
K_P
K_I
$G_{RE}(z)$
$\beta_D \cdot T$
325
200
322

Fig. 5B

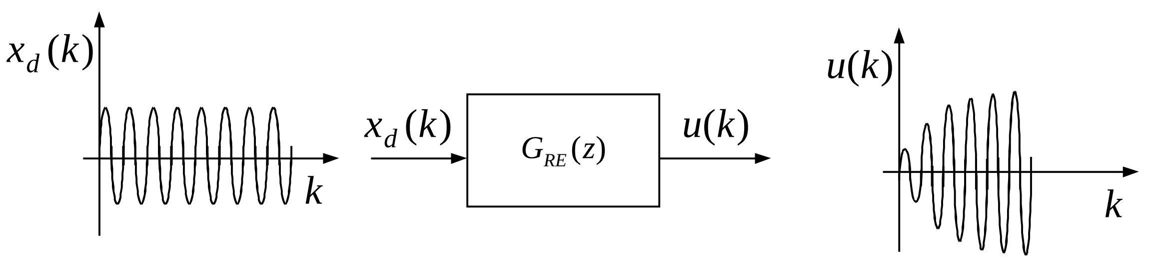

$x_d(k)$
k
$x_d(k)$
$G_{RE}(z)$
$u(k)$
$u(k)$
k

Fig. 6A

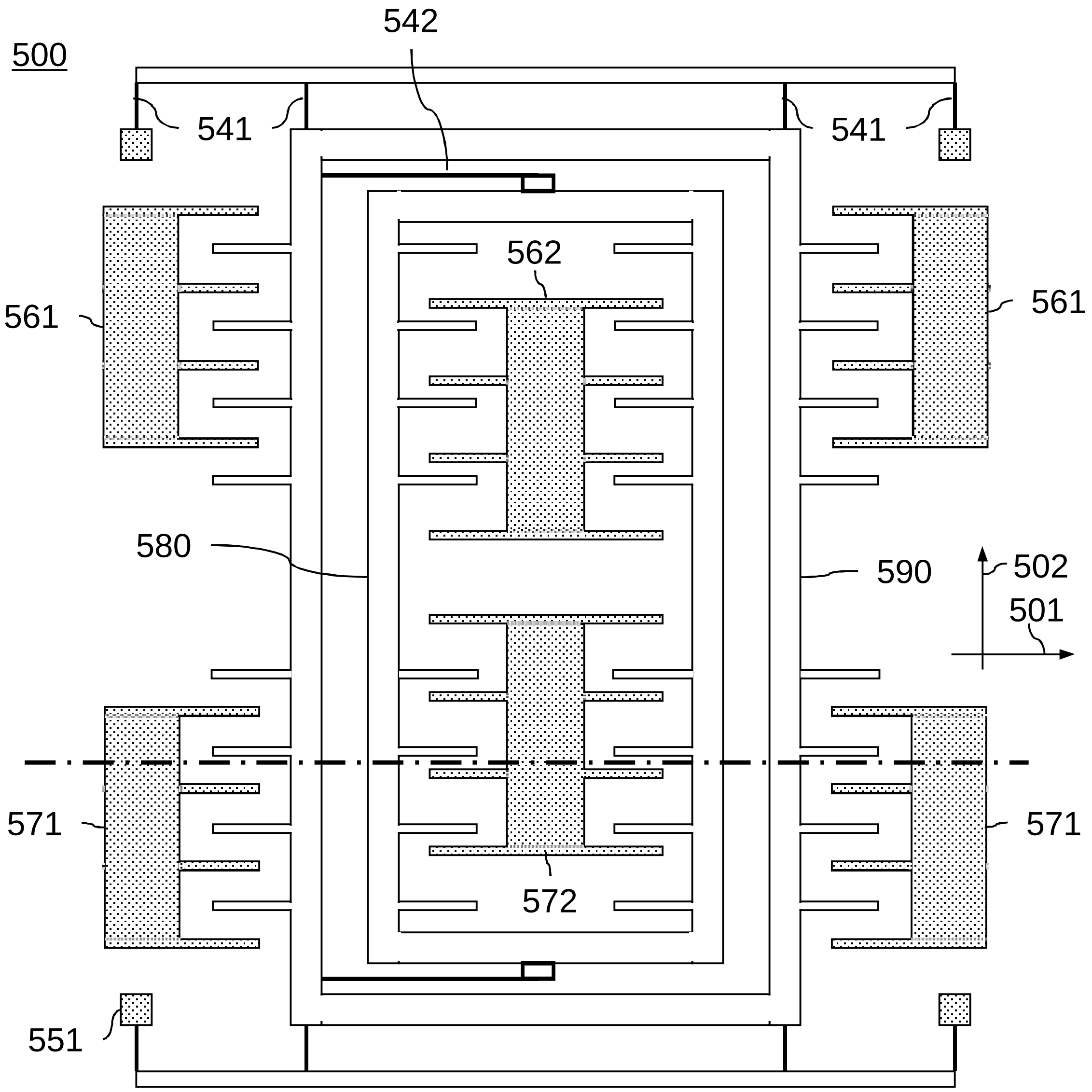

542
500
541
541
562
561
561
580
590
502
501
571
571
572
551

Fig. 6B

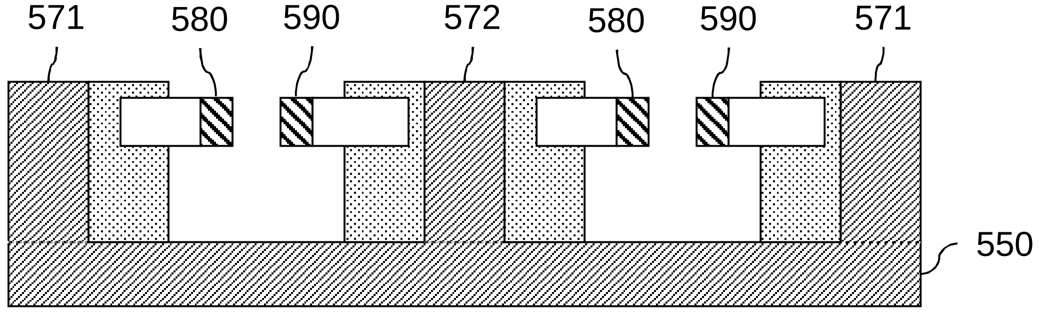

571
580
590
572
580
590
571
550

Fig. 6C

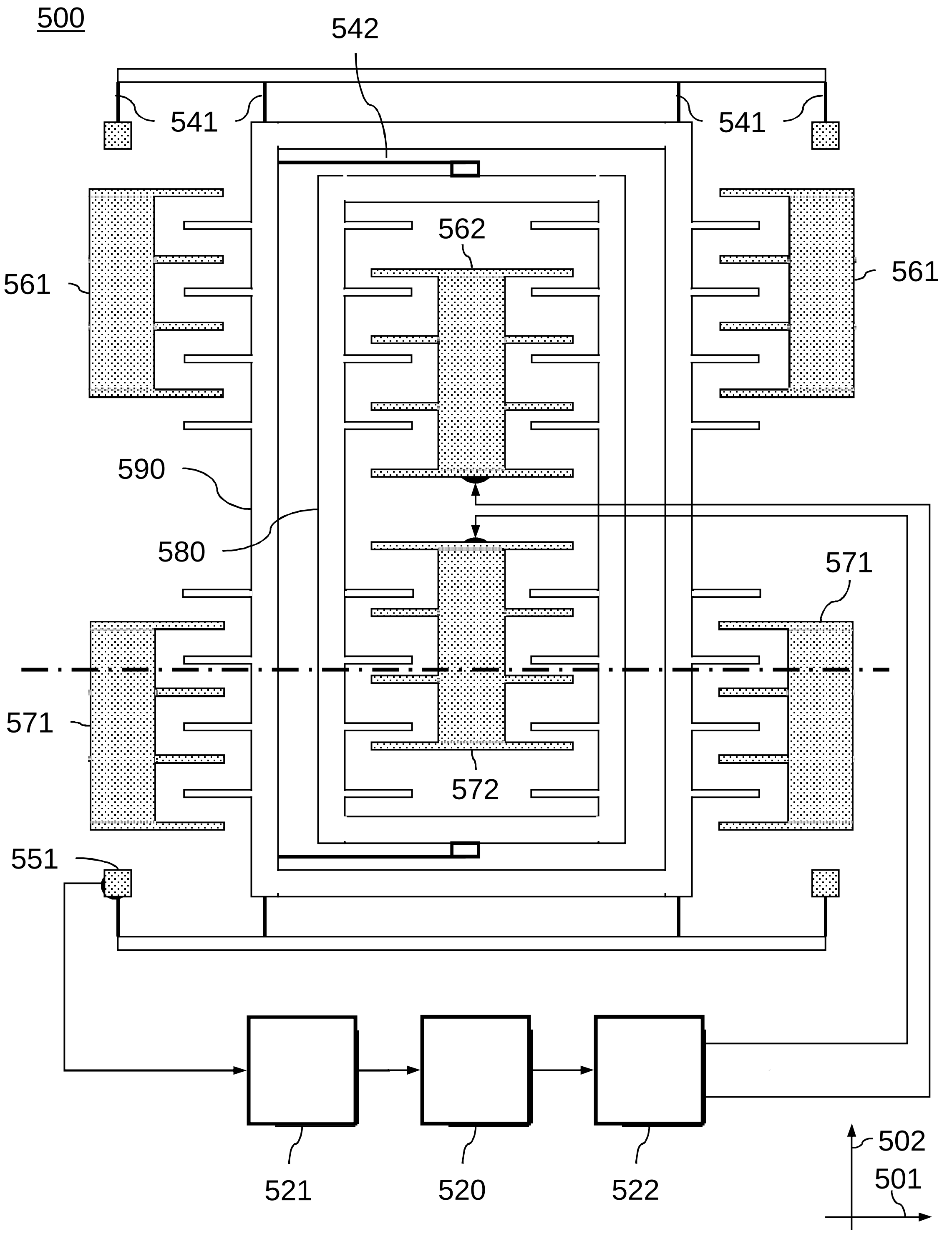

500
542
541
541
562
561
561
590
580
571
571
572
551
521
520
522
502
501

Fig. 7

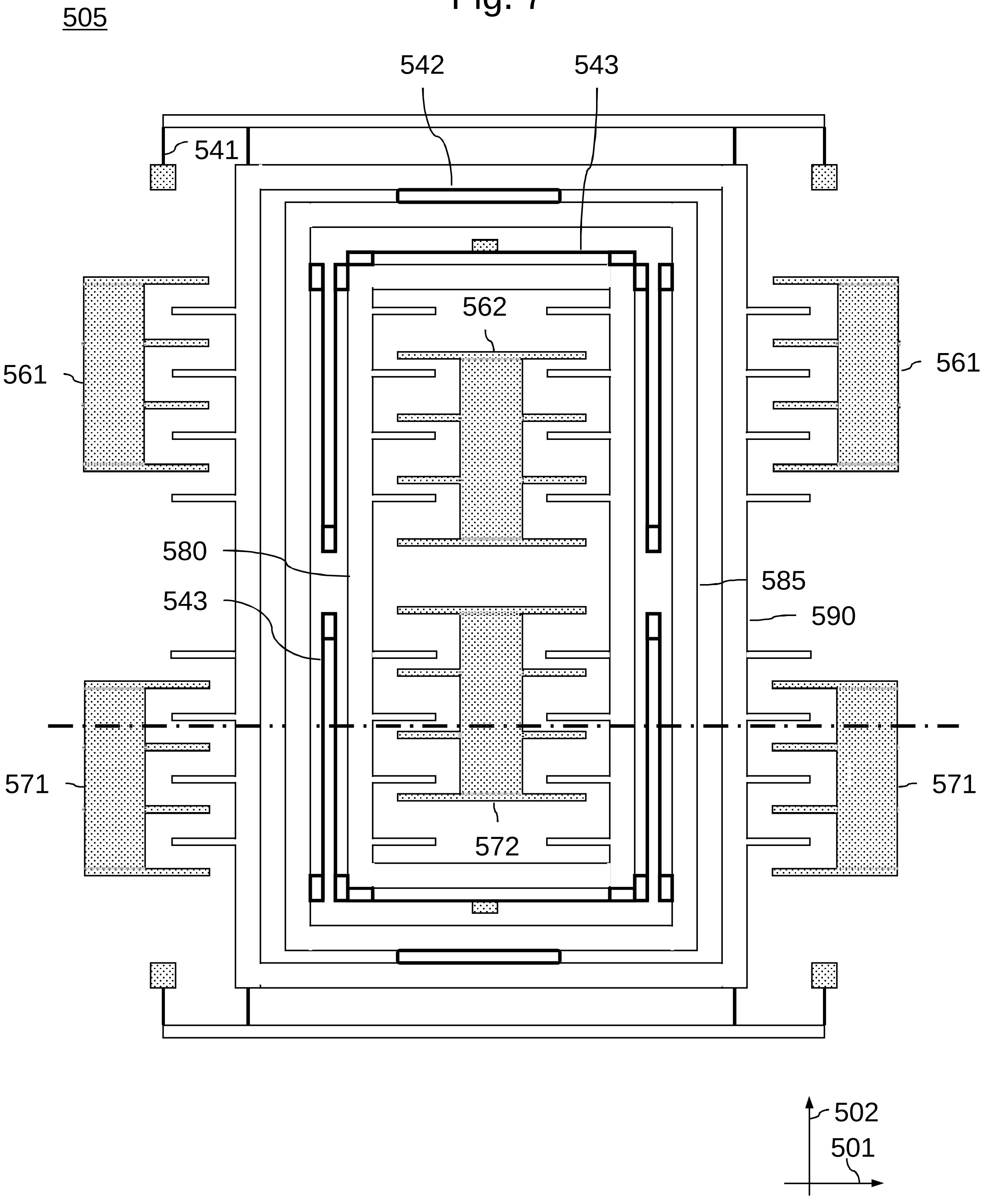
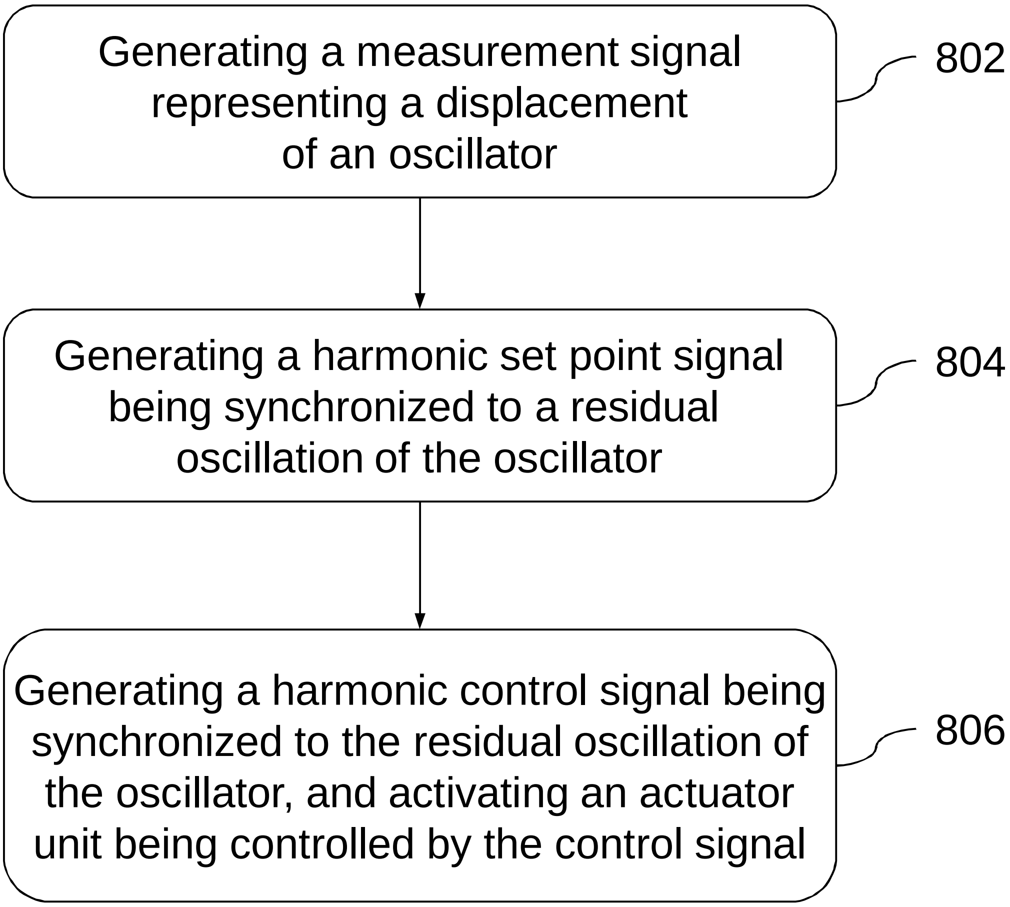


Fig. 8


Generating a measurement signal representing a displacement of an oscillator
802
Generating a harmonic set point signal being synchronized to a residual oscillation of the oscillator
804
Generating a harmonic control signal being synchronized to the residual oscillation of the oscillator, and activating an actuator unit being controlled by the control signal
806