Self-Tuning Algorithm for an Optical Pointing System with Variable Payloads

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Abstract: A self-tuning controller for a pointing system with variable mass/inertia payloads is presented. The self tuning algorithm is based on the on-line reaction curve of a change in voltage to current of the permanent magnet Direct Current (DC) motors. Adaptive control, sine testing and time-series methods are not considered due to physical system limitations and memory and processing limitations of the selected micro-processors. A classic cascaded control structure is used to implement the current, rate and position loops in that order and these are designed automatically and on-line based on a parameter set determined from the reaction curve. Both set-point changes and disturbances in high wind, with varied payloads, are easily dealt with by the system.

Keywords: pointing system, pan-and-tilt unit, self-tuning control, cascaded control, reaction curve, motor control, optical payload

1. INTRODUCTION

Optical trackers are necessary equipment for surveillance in law enforcement and military applications. The Stand Alone Tracker (SAT) used by Optronics Sensor Systems (OSS) at the Defence, Peace, Safety and Security (DPSS) unit of the Council for Scientific and Industrial Research (CSIR) has been revamped to include easier access for debugging purposes, robustness and a modular design so that several people can design and test their boards individually. The benefit of a modular design is that relative complex algorithms can be implemented on each of the modules and changes are easily carried out. The implemented control system is self-tuning since variable mass/inertia payloads are placed on the device.

SAT (see Fig. 1) consists of individually powered azimuth and elevation axes; with inductosyn resolver and optical encoder based position measurement, tacho-generator based rate measurement and Hall effect based current measurement on the permanent magnet Direct Current (DC) motors. The modularity of the system is achieved by creating modules for each of the main functions of the system and for each axis. These consist of a master-, rate-, position- and motor-controller (driver) modules, each with a Texas Instruments C2000 series control card (TMS320F28335 Texas Instruments (2007)) as the main Digital Signal Controller (DSC). The master modules control the events and are responsible for RS232 based supervisory control and rate or position commands on the RS485 port from external commanding systems such as radar trackers. Information and command transfer is achieved by means of the MCBSP (multichannel buffered serial port) communication interface running at 1kHz, which is controlled by the master module. Supervisory control, commanding and power is achieved through a copper slipring.

The control system of the pointing system resides entirely on the driver modules and accurate position and rate feedback is communicated through the MCBSP protocol and used by the driver module for control in a classic cascaded fashion: an inner loop which measures current and sets voltage PWM current/torque control, an intermediate loop which measures rate and sets the current set-point (rate control) and an outer loop which measures position and gives rate set-points. The self-tuning algorithm also resides on the driver module.

The typical elevation payload that is carried by SAT includes four cameras: Carl Zeiss Optronics Attica Z (3-5µm) and Attica P (8-12µm) cameras, OSS developed Cyclone (2.4× zoom long range surveillance camera with near infrared narrow Field-of-View (FOV) day channel with a concurrent wide FOV colour channel and separate electron multiplying charged coupled device based low light channel) and OSS developed visible/near infrared full high definition medium FOV camera (MFOVC) totalling 105kg in mass (including mountings). Several cameras may be interchanged from the system if needed and this is the purpose of the self-tuning controller. The payload data is transmitted through a capacitively coupled gigabit Ethernet slip-ring (Schleifring) to the outside world.

The control requirement is based on the FOV of Cyclone (0.19°). The requirement is positioning within 0.05° and
stabilization with typical coastal winds acting on the system. Furthermore, the system is not to exceed 1 rad/s.

This paper presents the self-tuning controller that was used to control both the azimuth and elevation axes.

Fig. 1. Optroinc Sensor Systems Stand Alone Tracker

2. DYNAMIC MODEL

The model describing any of the two axes (assuming that the azimuth and elevation axes are adequately decoupled which is the case here when the loads are balanced such that the azimuth and elevation rotation axes and their inertial axes coincide, respectively) is given by

\[
\begin{aligned}
\frac{di(t)}{dt} &= \begin{bmatrix} R & K & 0 \\ -\frac{1}{J} & \frac{1}{B} & 0 \\ 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} i(t) \\ \omega(t) \\ \theta(t) \end{bmatrix} + \begin{bmatrix} 1 \\ 0 \\ 0 \end{bmatrix} V(t) + \\
\frac{di(t)}{d\omega(t)} &= 0 = \begin{bmatrix} 0 \\ 0 \\ 0 \end{bmatrix} T_L(t),
\end{aligned}
\]

where

\[
\Delta(s) = Ls^2 + (LB + RJ)s + RB + K^2,
\]

is the characteristic equation of the system. \( L \) [H] is the inductance of the motor, \( J \) [kgm²] is the inertia seen by the motor, \( B \) [Nms/rad] is the viscous friction, \( R \) [Ω] is the resistance of the motor and \( K \) [Nm/A] is the torque constant. \( i \) [A], \( \omega \) [rad/s] and \( \theta \) [rad] are the current, angular rate and angular position of the axis in question. \( V \) [V] is the voltage and \( T_L \) [Nm] is the load torque (disturbance). Note that \( V(t) \leftrightarrow V(s) \) is the first input and \( T_L(t) \leftrightarrow T_L(s) \) is the second input to the system. This linear model is assumed to represent the system adequately for controller design and is used in the following for self-tuning. See Lyshevski (1999) for more detailed models.

3. PROCESS REACTION CURVE

Conventional methods use e.g. Box-Jenkins based models (ARX, ARMAX etc. see Ljung (1999)) or neural nets which require some form of large matrix inversions and non-parametric methods such as sine wave testing or self-tuning methods such as Model Reference Adaptive Control Koksal et al. (2007). These methods were not applicable to the current project since they either require a large memory space or powerful processors, take too long or are inherently bad for the system (e.g. high frequency oscillation may cause lenses and mechanical parts to become dislodged). The idea of the self-tuning system is to find the Process Reaction Curve (PRC or step response, Stephanopoulos (1984), Seborg et al. (1989)) of the current in the motor together with the steady rate during the step and use this information to design a controller on-line (see Seebacher et al. (2007) for an off-line method).

The step applied to the voltage of the motor using the PWM controller with no loaded torque applied except for the load itself results in the following theoretical step response (PRC) in the current (35% positive (bipolar) PWM was used with 36V supply voltage for this project):

\[
i(t) = c_1 + c_2 e^{c_3 t} + c_4 e^{c_5 t}
\]

\[
= \text{offset} + \text{slow dynamics} + \text{fast dynamics}.
\]

The units are [A/V], [A/V], [rad/s], [A/V] and [rad/s] for the coefficients \( c_1 \) to \( c_5 \) respectively. A typical response of \( i(t) \) is shown in Figure 2 (see Maamri et al. (2007) for a typical impulse response).

Fig. 2. Process Reaction Curve of Voltage to Current

Finding the coefficients from the real PRC is hard and a different approach (over SID, curve fitting, parameter estimation and iteration methods) was followed. Two methods are presented.

3.1 Method 1

The PRC has a natural peak at \( t_p \). \( t_p \) can be found from

\[
\frac{di(t_p)}{dt} = 0 = c_2 c_3 e^{c_3 t_p} + c_4 c_5 e^{c_5 t_p},
\]

and is given by

\[
t_p = \frac{1}{c_5 - c_3} \ln \left( \frac{-c_2 c_3}{c_4 c_5} \right).
\]
This equation is not used in the solution procedure as it is very difficult to solve together with the other equations (to be presented) and is not suitable for a microprocessor. c1 < 0 and c5 < 0 due to the implied stability of the system. The system is said to be in steady-state when the response has eventually reached 98% of its final value and the time at which this occurs is denoted by ts (same as settling time). At steady-state \((t \rightarrow \infty)\), the modes disappear and the current is the offset current:

\[ i(\infty) = c_1 = i_{s\infty}. \] (6)

Also, the current is zero and the modes are not acting yet when \(t = 0\) so that

\[ i(0) = 0 = c_1 + c_2 + c_4. \] (7)

Let \(c_3\) be associated with the time constant of the mechanical component \((\tau_m = J/B = -1/c_3 [s])\) of the system and \(c_5\) with the time constant of the electrical component \((\tau_e = L/R = -1/c_5)\) of the system. Generally, \(\tau_m \gg \tau_e\) (over-damped system), implying \(c_3 \gg c_5\) (normally \(c_3 \geq 0.01c_5\)). For this reason, the fast dynamic dominates the response while \(t\) is small; and the slow dynamic dominates the response when \(t\) is large but not at steady-state. Therefore, \(e^{c_5 t} \approx 1\) while \(t\) is small and the dynamic is dominated by the \(e^{c_5 t}\) term. Also, \(e^{c_5 t} \approx 0\) (the fast mode influence is negligible) while \(t\) is past the peak at \(t_p\). The dynamic is therefore dominated by the \(e^{c_5 t}\) term until its influence becomes negligible and the offset becomes the dominant dynamic. This implies that

\[ i(t) \approx c_1 + c_2 + c_4 e^{c_5 t} \quad \forall \quad 0 \leq t \leq t_p, \] (8a)

\[ i(t) \approx c_1 + c_2 e^{c_5 t} \quad \forall \quad t_p \leq t \leq t_{ss}. \] (8b)

Let the times within these periods be denoted by \(t_f\) and \(t_s\) i.e. \(0 \leq t_f \leq t_p\) and \(t_p \leq t_s \leq t_{ss}\). Also let \(i(t_f) = i_f\) and \(i(t_s) = i_s\). Furthermore,

\[ \frac{di(t)}{dt} \approx c_4 c_5 e^{c_5 t} \quad \forall \quad 0 \leq t \leq t_p, \] (9a)

\[ \frac{di(t)}{dt} \approx c_2 c_3 e^{c_5 t} \quad \forall \quad t_p \leq t \leq t_{ss}. \] (9b)

Let \(\frac{di(t_f)}{dt} = i'_f\) and \(\frac{di(t_s)}{dt} = i'_s\). There are now six equations (6, 7, 8 and 9) with five unknowns \((c_1\) to \(c_5\)). One equation can therefore be eliminated and equation 8a is eliminated because it makes the solution of the unknowns simpler than using other sets of equations (trial and error based on testing data—not shown here). Testing the solutions of the coefficients in equation 8a, however, showed that the solutions held for this equation as well. Note that \(i_{s\infty}, i_s, i'_s, i'_f\) are to be determined from the PRC. \(t_s\) should be selected somewhere between \(t_p\) and \(t_{ss}\); the midway point can be used \((t_s = 0.5[t_{ss} - t_p])\). A summary of the equations are given as (with \(t_f\) selected at 0):

\[ i_{s\infty} = c_1 \] \hspace{1cm} (10a)

\[ i'_s = c_2 c_3 e^{c_5 t_s} \] \hspace{1cm} (10b)

\[ i_s = c_1 + c_2 e^{c_5 t_s} \] \hspace{1cm} (10c)

\[ 0 = c_1 + c_2 + c_4 \] \hspace{1cm} (10d)

\[ i'_f = c_4 c_5. \] \hspace{1cm} (10e)

Solution of the unknowns gives

\[ 0 = -i'_f - c_5 \left[ i_{s\infty} + (i_s - i_{s\infty}) \exp \left( \frac{i'_f t_s}{i_{s\infty} - i_s} \right) \right] \] (11a)

\[ 0 = i_{s\infty} + (i_s - i_{s\infty}) \exp \left( \frac{i'_f t_s}{i_{s\infty} - i_s} \right) + c_4 \] (11b)

\[ 0 = -i_s + i_{s\infty} + \frac{i'_f}{c_5} \] (11c)

\[ 0 = -i'_s + c_2 c_3 e^{c_5 t_s} \] (11d)

\[ 0 = -i_{s\infty} + c_1, \] (11e)

with further manipulation gives the coefficients:

\[ c_5 = \frac{i'_f}{-i_{s\infty} - (i_s - i_{s\infty}) \exp \left( \frac{i'_f t_s}{i_{s\infty} - i_s} \right)} \] (12a)

\[ c_4 = -i_{s\infty} - (i_s - i_{s\infty}) \exp \left( \frac{i'_f t_s}{i_{s\infty} - i_s} \right) \] (12b)

\[ c_3 = \frac{i'_f}{i_{s\infty} - i_s} \] (12c)

\[ c_2 = (i_s - i_{s\infty}) \exp \left( \frac{i'_f t_s}{i_{s\infty} - i_s} \right) \] (12d)

\[ c_1 = i_{s\infty}. \] (12e)

This method proved valuable during initial testing but was not repeatable enough as variations of up to 20% were found in the coefficients on subsequent self-tuning experiments—another approach was needed.

### 3.2 Method 2

An alternative method is presented here. Consider equation 3 again. At steady-state, the fast and slow dynamics have died out and

\[ c_1 \rightarrow i_{s\infty} \forall \quad t \rightarrow t_{\infty}. \] (13)

Before steady-state, the slow dynamic is dominant and the fast dynamic effect is very small so that the step response is governed by

\[ i_s(t) \approx c_1 + c_2 e^{c_5 t} \forall \quad t_p < t < t_{ss}. \] (14)

Using step data and the following transformation one can determine \(c_2\) and \(c_3\) by regression (see e.g. Mathews (1992)):

\[ i_s(t) \approx c_1 + c_2 e^{c_5 t} \] (15)

\[ \therefore \ln \left[ i_s(t) - c_1 \right] \approx \ln \left[ c_2 e^{c_5 t} \right] \] (16)

\[ = c_3 t + \ln[c_2] \] (17)

\[ = \hat{m}_s t + \hat{c}_s. \] (18)

The transformation results in a straight line approximation with a slope \((\hat{m}_s)\) and an intercept \((\hat{c}_s)\). These can be determined by linear regression. Once the slope and intercept are known, \(c_2\) and \(c_3\) can be determined from

\[ c_2 = e^{\hat{c}_s} \] (19)

\[ c_3 = \hat{m}_s. \] (20)

The data set used for the slow dynamic determination in this project was \(2t_p < t < 12t_p\), which delivered the best results. Once the slow response parameters are known, the fast response parameters can be determined as follows: since estimates of \(c_1\), \(c_2\) and \(c_3\) are known, the
slow response is subtracted from the total response and \( c_4 \) and \( c_5 \) are determined through regression in a similar way as above:

\[
i(t) \approx c_1 + c_2 e^{ct} + c_4 e^{cs} \]

\[
\therefore -i(t) + c_1 + c_2 e^{ct} \approx -c_4 e^{cs} \tag{22}
\]

\[
\therefore \ln \left[ -i(t) + c_1 + c_2 e^{ct} \right] \approx \ln \left[ -c_4 e^{cs} \right] \tag{23}
\]

\[
= c_5 t + \ln \left[ -c_4 \right] \tag{24}
\]

\[
= m_i t + \hat{c}_f. \tag{25}
\]

The data set from \( 0 < t < 0.9 t_p \) was used as this produced the best results. This negative sign is necessary because \( c_4 < 0 \). After regression, \( c_4 \) and \( c_5 \) can be determined from

\[
c_4 = -e^{ct} \tag{26}
\]

\[
c_5 = \hat{m}_i. \tag{27}
\]

With the parameters known, controllers can be designed for the system on-line.

4. CONTROLLER

The (classic) controller structure is presented in figure 3. The idea for this form of control is to eliminate the inclusion of a dominant zero in the current response when rate or position control is applied directly using the voltage as the input. This causes less saturation of the current and thereby one can achieve a better bandwidth of the system without ringing.

Control starts in the inner current or loop since the response is the fastest here. The rate loop is the middle loop and if position control is desired, the position loop becomes the outer loop. To limit the effects of saturation of the current and rate, the middle and outer (rate and position) loop gains can be de-tuned. The (somewhat conservative) control objective was for the system to respond in a critically damped way with about 80\% saturation over time on all the signals present. The saturation on the position and rate occur at the set-points as it is assumed that the system will be critically damped on all variables and thus these variables will not go over the saturation value (no direct mechanism exists to limit current, rate or position). Note that the controller is automatically designed on-line, using the determined coefficients.

5. CURRENT CONTROLLER

Now, assuming that a step with a magnitude of \( A \) volts was applied in the voltage, then the transfer function of the voltage to current system is

\[
g_{i,v}(s) = \frac{(-c_1 c_5 - c_1 c_3 - c_2 c_5 - c_4 c_3) s + c_1 c_3 c_5}{A s^2 + A(-c_3 - c_5) s + A c_3 c_5} \tag{28}
\]

where the coefficients are given by

\[
b_1 = (-c_1 c_5 - c_1 c_3 - c_2 c_5 - c_4 c_3)
\]

\[
b_0 = c_1 c_3 c_5
\]

\[
a_2 = A
\]

\[
a_1 = A(-c_3 - c_5)
\]

\[
a_0 = A c_3 c_5, \tag{29}
\]

and the steady-state gain is

\[
k = \frac{c_1}{A}. \tag{30}
\]

Since the system is stable and minimum phase, a controller which inverts the plant is sufficient if integration is added. The controller is therefore chosen as

\[
c_i(s) = \frac{K_i}{s} g_{i,v}(s) = \frac{K_i}{s} \frac{a_2 s^2 + a_1 s + a_0}{b_1 s + b_0}, \tag{31}
\]

where \( K_i \) is the gain (ultimately the bandwidth of the closed-loop system, which can be selected to be more than the corner frequency of the electrical time constant (\( K_i \geq \frac{1}{\tau_e} = -c_5 \)). A value of \( K_i = 1/\tau_e = -c_5 \) was used in this project. The closed-loop transfer function is

\[
T_{i,i,p}(s) = \frac{1}{K_i s + 1}. \tag{32}
\]

The digital version of the controller can be found through Tustin’s transform by setting \( T \) is the sampling and control period

\[
s = \frac{2 z - 1}{T z + 1}. \tag{33}
\]

The \( z \)-transformed transfer function is given by

\[
c_i(z) = \left[ (4 a_2 + 2 a_1 T + a_0 T^2) K_i + (-8 a_2 + 2 a_0 T^2) K_i \right. \\
\]

\[
+ (4 a_2 - 2 a_1 T + a_0 T^2) K_i, \tag{34}
\]

\[
\left. z^{-1} \right]
\]

\[
+ \frac{(4 b_1 + 2 b_0 T) - 8 b_1 z^{-1} + (4 b_1 - 2 b_0 T) z^{-2}}{1 - 8 b_1 z^{-1} + (4 b_1 - 2 b_0 T) z^{-2}},
\]

and can be written as

\[
c_i(z) = \frac{v_0 + w_1 z^{-1} + w_2 z^{-2}}{v_0 + v_1 z^{-1} + v_2 z^{-2}}. \tag{35}
\]

This can further be written as

\[
v_0 V(z) = -v_1 V[z - 1] - v_2 V[z - 2]
\]

\[
+ w_0 E_i[z] + w_1 E_i[z - 1] + w_2 E_i[z - 2], \tag{36}
\]

where \( V[z] \) is the voltage and \( E_i(s) \) is the current error and

\[
w_0 = (4 a_2 + 2 a_1 T + a_0 T^2) K_i, \quad v_0 = (4 b_1 + 2 b_0 T)
\]

\[
w_1 = (-8 a_2 + 2 a_0 T^2) K_i, \quad v_1 = -8 b_1
\]

\[
w_2 = (4 a_2 - 2 a_1 T + a_0 T^2) K_i, \quad v_2 = (4 b_1 - 2 b_0 T). \tag{37}
\]

6. RATE CONTROLLER

Assuming that the estimate of the voltage to current controller was perfect, then the controller will invert the voltage to current plant as previously discussed. The transfer function from the voltage to the rate is given by

\[
g_{\omega,v}(s) = \frac{b_2}{a_2 s^2 + a_1 s + a_0}, \tag{38}
\]

where \( b_2 \) is found from the steady-state value of the rate in the test in \( g_3. \) Specifically,

\[
b_2 = a_0 \omega_{ss}, \tag{39}
\]

where \( \omega_{ss} \) is the steady-state value of the rate in the test in \( g_3. \) \( a_2, a_1, \) and \( a_0 \) are the same as in \( g_3. \) The transfer function of the current to rate with the current controller implemented is

\[
g_{\omega,i,p}(s) = \frac{K_i b_2}{(s + K_i) (b_1 s + b_0)} = \frac{K_i b_2}{b_1 \left( \frac{1}{s + K_i} \right) \left( s + \frac{b_0}{b_1} \right)}, \tag{40}
\]
Fig. 3. Controller Block Diagram

A proportional controller is sufficient for the large signal movement of the system (i.e., \( c_\omega(s) = K_\omega \)). The controller gain can be chosen to achieve a critically damped response (the two poles of the current to rate transfer function with current control gives two poles). The gain chosen (from Root-Locus theory) as

\[
K_\omega = \frac{(K_1 b_1 - b_0)^2 (2 - d_\omega)}{4K_1 b_0 b_\omega},
\]

where \( 0 < d_\omega \leq 1 \) gives the location of the closed-loop pole compared to the critically damped case (when \( d_\omega = 1 \), the closed-loop poles are situated at \( \frac{1}{2} (K_1 + b_0 \theta_1) \) and when \( d_\omega = 0 \) the closed-loop poles are at the open-loop poles). A value of \( d_\omega = 1 \) was used to reduce the effects of current saturation.

Alternatively, a PI controller can be used (especially due to the dead band at low current) of the form

\[
c_\omega(s) = K_\omega \frac{s + b_0 \theta_1}{s},
\]

(43)
to cancel out the slow pole and add integration. The gain is then selected as

\[
K_\omega = \frac{K_1 b_1 (2 - d_\omega)}{4b_0 \omega},
\]

(44)
This controller is discretized as

\[
c_\omega[z] = \frac{K_\omega (2b_1 + b_0 T) + K_\omega (-2b_1 + b_0 T) z^{-1}}{2b_1 - 2b_0 T z^{-1}}.
\]

(45)
This can be written as

\[
c_\omega[z] = \frac{p_0 + p_1 z^{-1}}{q_0 + q_1 z^{-1}},
\]

(46)
with

\[
p_0 = K_\omega (2b_1 + b_0 T) \quad q_0 = 2b_1
\]

(47)
\[
p_1 = K_\omega (-2b_1 + b_0 T) \quad q_1 = -2b_1.
\]

7. POSITION CONTROLLER

The closed-loop transfer function from the rate set-point, \( \omega_{\theta,p} \), to the rate is given by

\[
T_{\omega,\omega_p}(s) = \frac{c_\omega(s)K_\omega b_\omega}{c_\omega(s)K_\theta b_\omega} = \frac{1}{\frac{c_\omega(s)K_\omega b_\omega}{b_1}} \frac{1}{s + K_\omega \left( s + \frac{b_0}{b_1} \right)} + 1
\]

(48)
For the case where proportional control was used for the rate loop,

\[
T_{\omega,\omega_p}(s) = \frac{1}{1 + \frac{b_1}{K_\omega K_\omega b_\omega} \left( s + K_\omega \right) \left( s + \frac{b_0}{b_1} \right)},
\]

(49)
and

\[
g_{\omega,\omega_p}(s) = \frac{1}{s + \frac{b_1}{K_\omega K_\omega b_\omega} \left( s + \frac{b_0}{b_1} \right)},
\]

(50)
and for the case where the PI controller was used,

\[
T_{\omega,\omega_p}(s) = \frac{1}{s + \frac{b_1}{K_\omega K_\omega b_\omega} s^2 + \frac{b_0}{b_1} s + 1},
\]

(51)
and

\[
g_{\omega,\omega_p}(s) = \frac{1}{s + \frac{b_1}{K_\omega K_\omega b_\omega} s^2 + \frac{b_0}{b_1} s + 1},
\]

(52)
Since integral action is already present for both types of rate controllers, proportional control will suffice (\( c_\theta(s) = K_\theta \)), and the closed-loop for the case where proportional rate control was used is given by

\[
T_{\theta,\omega_p} = \frac{1}{s + \frac{b_1}{K_\theta b_\omega} \left( s + \frac{b_0}{b_1} \right)},
\]

(53)
\[= \frac{1}{s + \frac{b_1}{K_\theta b_\omega}} \left( s + \frac{b_0}{b_1} \right) + 1
\]
\[
= \frac{1}{s + \frac{b_1}{K_\theta b_\omega}} \left( s + \frac{b_0}{b_1} \right)
\]
\[
= \frac{1}{s + \frac{b_1}{K_\theta b_\omega}} \left( s + \frac{b_0}{b_1} \right)
\]
\[
= \frac{1}{s + \frac{b_1}{K_\theta b_\omega}} \left( s + \frac{b_0}{b_1} \right)
\]
\[
= \frac{1}{s + \frac{b_1}{K_\theta b_\omega}} \left( s + \frac{b_0}{b_1} \right)
\]
\[
= \frac{1}{s + \frac{b_1}{K_\theta b_\omega}} \left( s + \frac{b_0}{b_1} \right)
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= \frac{1}{s + \frac{b_1}{K_\theta b_\omega}} \left( s + \frac{b_0}{b_1} \right)
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= \frac{1}{s + \frac{b_1}{K_\theta b_\omega}} \left( s + \frac{b_0}{b_1} \right)
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= \frac{1}{s + \frac{b_1}{K_\theta b_\omega}} \left( s + \frac{b_0}{b_1} \right)
\]
\[
= \frac{1}{s + \frac{b_1}{K_\theta b_\omega}} \left( s + \frac{b_0}{b_1} \right)
\]
\[
= \frac{1}{s + \frac{b_1}{K_\theta b_\omega}} \left( s + \frac{b_0}{b_1} \right)
\]
Equating equivalent terms and substituting for known parameters and \( K_\theta \) in \( \alpha_3 \) gives the gain for critical damping:

\[
K_\theta = \frac{b_1}{K_\theta b_\omega} \left( \frac{1}{3} \alpha_1 \alpha_2 - \frac{2}{27} \alpha_3 \right) + \frac{2}{27} \sqrt{\left( \alpha_1^2 - 3 \alpha_2 \right)^3}.
\]

(56)
The controller should be tuned by a factor of \( d_\theta \) (two fifths (2/5) was used in this project) to reduce the effects of rate saturation. This roughly implies a position bandwidth equal to one tenth of the bandwidth of the rate loop.

For the case where a PI controller was used for the rate loop, \( b_0 \) is simply made zero and (55) becomes

\[
\alpha_1 = K_\theta
\]

(57a)
\[
\alpha_2 = \frac{K_\omega K_\theta b_\omega}{b_1}
\]

(57b)
\[
\alpha_3 = \frac{K_\omega K_\theta b_\theta}{b_1}
\]

(57c)
8. RESULTS

The system was tested in Simon’s Town (near Cape Town) at the coast on 13 October 2010 with a payload of Cyclone and MFOVC on one occasion; and additionally the two Attica cameras on a separate occasion (27 September 2010). The presented current controller was used and the PI controller was used for the rate controller and the associated proportional controller for position control. Method 2 was used to determine the unknown parameters. All loops were set to ensure that the responses were either critically- or over-damped but never under-damped by manipulating $K_i$, $d_\omega$ and $d_\theta$. For the azimuth (left) and elevation (right) axes, the parameters in table 1 were determined automatically and on-line for the case of Cyclone and MFOVC (13 October 2010). An azimuth step change is shown in Fig. 4 and associated elevation response in Fig. 5. The positioning errors are 0.03° and 0.04° for the azimuth- and elevation-axes respectively, as required. Furthermore, the speed specification (<1rad/s) is not exceeded.

The self-tuning occurred flawlessly (in low wind) and the controller tested well in typical Simon’s Town winds of 30 knots.

Table 1. Results of the Test for Two Cameras

<table>
<thead>
<tr>
<th></th>
<th>Azimuth</th>
<th>Elevation</th>
</tr>
</thead>
<tbody>
<tr>
<td>$c_1$</td>
<td>0.309</td>
<td>0.151</td>
</tr>
<tr>
<td>$c_2$</td>
<td>0.497</td>
<td>1.553</td>
</tr>
<tr>
<td>$c_3$</td>
<td>-3.873</td>
<td>-39.61</td>
</tr>
<tr>
<td>$c_4$</td>
<td>-0.611</td>
<td>-1.644</td>
</tr>
<tr>
<td>$c_5$</td>
<td>-352.8</td>
<td>-225.4</td>
</tr>
<tr>
<td>$\omega_\phi$</td>
<td>0.448</td>
<td>0.998</td>
</tr>
<tr>
<td>$c_{\omega_\phi}$</td>
<td>46.054</td>
<td>0.161</td>
</tr>
<tr>
<td>$c_{\theta}$</td>
<td>46.054</td>
<td>6.688</td>
</tr>
</tbody>
</table>

Fig. 4. Azimuth Response

Fig. 5. Elevation Response

9. CONCLUSION

A method for self-tuning of an azimuth and elevation pointing system with variable optical loads using a microcontroller system was presented. Self-tuning was based on the reaction curve of the PWM voltage to the motors. Linear regression of the step response of the system was used to determine the unknown $c_1 - c_5$ parameters and steady rate on-line. The controllers were set automatically on-line based on the result of the reaction curve parameters. A classic structure for control was used where current, rate and position loops were closed automatically in that order and to deliver critically damped responses. Testing of the system delivered excellent results with disturbances in the range of 30 knots. Test and refinement is ongoing.

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REFERENCES


