

Article

Design and Implementation of Attitude Stabilization System for Marine Satellite Tracking Antenna

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Abstract: Marine Satellite Tracking Antenna (MSTA) is an important shipboard device for ships to bidirectionally communicate with the outside. The attitude stabilization system is an important part of MSTA for keeping the antenna dish tracking the geostationary satellite in the presence of severe ship dynamics. In this paper, the designed high-performance attitude stabilization system is introduced, including hardware system, model identification and controller design. The detailed procedure for identifying the stepper motor model is stated, which is different from conventional procedure. The design procedure of the robust controller based on $\mathcal{H}_\infty$ loop shaping is given. The designed attitude stabilization system was tested on a Stewart platform that was used as the ship simulator. From the testing results, it can be seen that the performance of designed attitude stabilization system is very high and the tracking angle error can be limited to within ± 0.15 deg, which satisfies the performance requirements of Ka-band MSTA product.

Keywords: attitude stabilization; marine satellite tracking antenna; robust control; $\mathcal{H}_\infty$ loop shaping

1. Introduction

Marine Satellite Tracking Antenna (MSTA) is an important shipboard device, which is responsible for establishing the communication between ships and geostationary satellites. The antenna dish of MSTA has to be directed at the geostationary satellite whatever the ship would rotate, due to ocean waves, as shown in Figure 1. To guarantee that high-quality satellite signal can be received by the antenna dish, the tracking angle error should be as small as possible. For the Ka-band MSTA under development (the latest technology in MSTA), the Root Mean Square (RMS) value of tracking angle error should be less than 0.2 deg under the condition that the maximum ship rotation velocity is about 26.2 deg/s. To realize this end, the design of attitude stabilization system of Ka-band MSTA, including algorithm, software and hardware, is very challenging.

Now, there are several different kinds of MSTA in the market, such as L-band, C-band and Ka-band MSTA, which are mainly divided according to the frequency range of received satellite signals. For example, the frequency range of the satellite signals received by the L-band and C-band MSTA is typically below 8 GHz, which is a relatively low frequency in field of satellite communication and also means that the tracking angle error of the antenna dish of MSTA can be relatively large, around 2 deg. However, the Ka-band MSTA, as the latest technology of MSTA, requires a high-accuracy tracking system. This is because the frequency range of the received satellite signals is from 26.5 GHz to 40 GHz, which enables greater volumes of data traffic to be transmitted. There is currently a great need for high bandwidth communication from those latest applications, such as high-definition television and broadband network, which means Ka-band MSTA has a massive market.

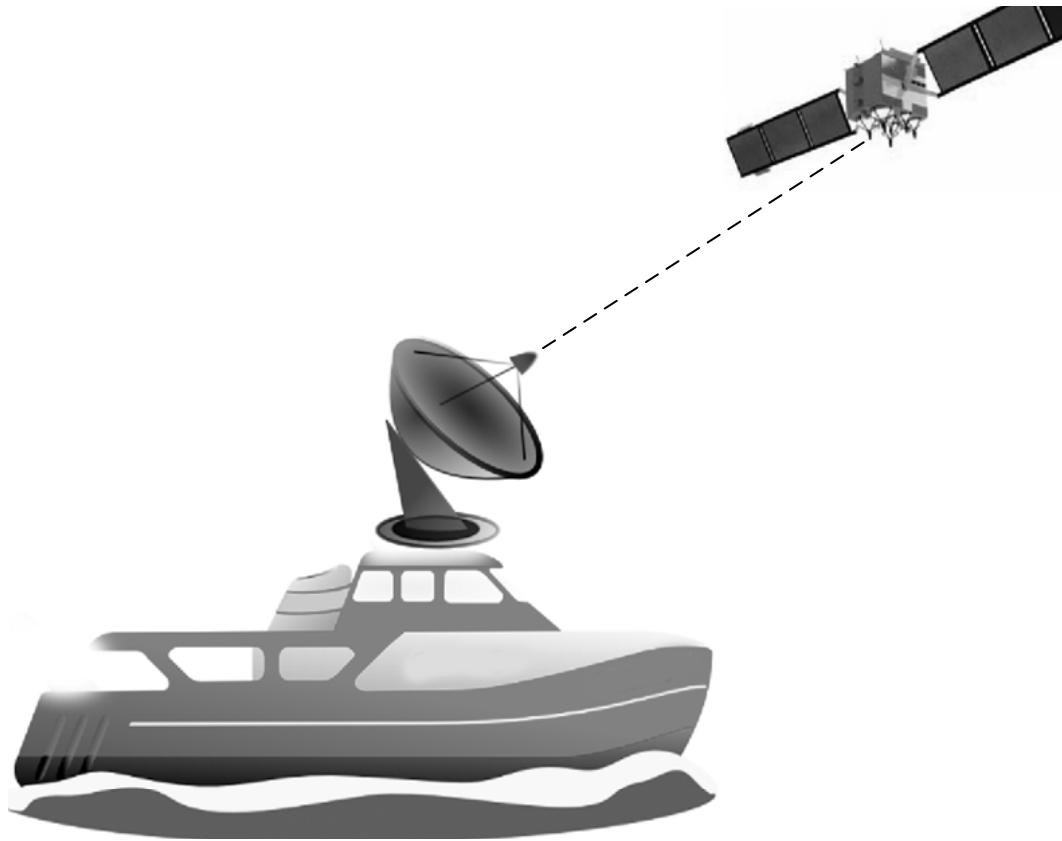


Figure 1. Illustration of MSTA receiving satellite signals.

Several research works have been done in the field of attitude control of MSTA. In [1], two generalized predictive control algorithms are used to control x and y axes of a two-degree-of-freedom robot, which is an earth station antenna related to be the HDF pedestals (High Dynamic Full Motion Leo Satellite Tracking Pedestals). In [2], a Fault Tolerant Control (FTC) system is proposed for the satellite tracking antenna. The FTC system maintains the tracking functionality by employing proper control strategy. A robust fault diagnosis system is designed to supervise the FTC system. The employed fault diagnosis solution is able to estimate the faults for a class of nonlinear systems acting under external disturbances. Fuzzy logic is used in [3] to cope with the uncertainty and nonlinear nature of the attitude control of shipboard tracking antenna. The simulations demonstrated the effectiveness of the design, which has been implemented on some antennas. In [4], a digital servo system of the Permanent Magnet Synchronous Motor (PMSM) is implemented in the satellite antenna position control. The novel control system based on Digital Signal Processing (DSP) and Complex Programmable Logic Device (CPLD) modules is proposed to satisfy the requirements of the high performance servo system. In [5], an attitude stabilization controller based on $\mathcal{H}_\infty$ mixed-sensitivity synthesis method is proposed to realize high accuracy of tracking control of MSTA. Experimental results show that the tracking accuracy can be improved to ± 0.3 deg, which is five times higher than that by the conventional controller. These research works have contributed a lot to the design of attitude stabilization system of MSTA, but there still exist some points that can be improved. The designed attitude control system of MSTA is fundamentally an attitude stabilization system, which is also applied in many other systems, besides MSTA system. In [6], two fault-tolerant control (FTC) schemes for spacecraft attitude stabilization with external disturbances are proposed. The effectiveness of the proposed schemes against actuator faults is demonstrated in simulation. In [7], the finite-time fault tolerant attitude stabilization control problem for a rigid spacecraft in the presence of actuator faults or failures, external disturbances, and modelling uncertainties is addressed. Again, only simulation studies are presented to demonstrate the effectiveness of the

proposed method. In these two papers, the proposed methods are not validated in real system, in which the working condition is much more complicated than simulation condition. In [8], a non-smooth attitude-stabilizing control strategy based on the finite-time control technique is investigated for flexible space crafts. In [9], a quaternion-based feedback control scheme for a quadrotor is proposed. The proposed controller is based upon the compensation of the Coriolis and gyroscopic torques and the use of a PD^2 feedback structure. The problems of attitude stabilization and disturbance torque attenuation for a small spacecraft using magnetic actuators is considered in [10]. In [11], an original configuration of a small aerial vehicle having eight rotors is presented. Four rotors are devoted to the stabilization of the orientation of the helicopter, and the other four are used to drive the lateral displacements. In [12], a fault-tolerant control approach without rate sensors is presented for the attitude stabilization of a satellite being developed. These works have contributed a lot to the attitude control of MSTA, but because of the differences of control purpose, actuators and attitude sensors, few of these research results can be applied on the control system of MSTA.

In this paper, the design and implementation of an attitude stabilization system of Ka-band MSTA is presented. The contributions of the present work are two-fold: (1) As the stepper motor is used in a close-loop system with the gyroscope for measuring shaft rotation velocity, the mathematical model of the stepper motor is identified with a method that is different from conventional method. (2) The robust controller based on $\mathcal{H}_\infty$ loop shaping method is stated, which can make the tracking error less than ± 0.15 deg under the condition that maximum ship rotation velocity can reach 26.2 deg/s.

The remainder of this paper is organized as follows: In Section 2, the diagram of the attitude stabilization system is introduced and two important electronic boards, the attitude sensor board and stepper motor control and drive board, are illustrated. In Section 3, the mathematical model of stepper motor is derived and some experiments are done to identify unknown parameters. The design procedure of a robust controller based on $\mathcal{H}_\infty$ loop shaping is stated in Section 4. The designed attitude stabilization system is tested in Section 5 on a Stewart platform and the experimental results are analyzed. Lastly, Section 6 gives the conclusion.

2. Introduction of Attitude Stabilization System

Attitude stabilization is an active research direction [12–15], and is also the core part of MSTA. The diagram of the attitude stabilization system is shown in this part. Two important self-designed boards are also introduced. The motivation for self designing boards is to explore those new and high-end electronic chips. These chips usually have great potential to increase the performance of the control system while there is no corresponding commercial boards on the market.

There are three parts in the MSTA under research, that is, the elevation part, cross-elevation part, and azimuth part. The controller design procedure for each part is almost the same and they are independent from each other. Thus, only the controller design of elevation part is selected for simplicity purpose.

2.1. Diagram of the Attitude Stabilization System

The diagram of the attitude stabilization system of elevation part is shown in Figure 2.

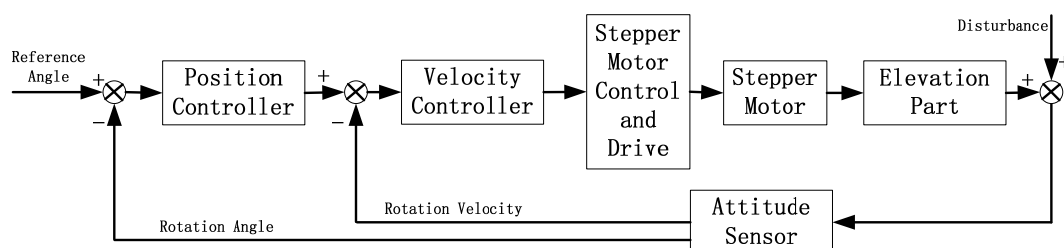


Figure 2. Diagram of the attitude stabilization system of elevation part.

In Figure 2, there are two control loops, the inner loop and the outer loop. The inner loop is the velocity loop, whose main function is to counteract the velocity disturbance from ships. Most of the disturbance is counteracted by the inner loop and the main focus is put on the design of the velocity controller. The velocity controller is designed based on $\mathcal{H}_\infty$ loop shaping. The disturbance that cannot be counteracted by the inner loop is left to the outer loop. The outer loop is the position loop and the position controller is a simple PI controller, whose design procedure is not stated.

2.2. Attitude Sensor Board

Attitude sensor is a key part in the attitude stabilization system, whose main function is to output the current rotation velocity and angle of the antenna dish. It is mounted together with the antenna dish and mainly consists of one Advanced RISC Machine (ARM) chip, and one Inertial Measurement Unit (IMU). The photo of the attitude sensor board is shown in Figure 3. The ARM chip (Cortex-M7 core) with the CPU frequency of 216 MHz reads sensor data from the IMU module through Serial Peripheral Interface (SPI) bus port. Inside the IMU module, there is one tri-axis gyroscope, one tri-axis accelerometer and one temperature sensor. The core parameters of the IMU module are listed in Table 1. The sensor fusion algorithm executed in the ARM chip fuses data from the IMU module and output the current Euler angles of the antenna dish. The popular attitude sensor fusion algorithm can be found in [16–19] and the definition of Euler is the same as that in [20]. The running results of the sensor fusion algorithm together with filtered gyroscope measurements are sent out through Universal Asynchronous Receiver/Transmitter (UART) communication port to the motor control and drive board with 1 Mbps baud rate.

Table 1. Core parameters of the IMU.

Sensor	Parameter	Value	Unit
Gyroscope	In-Run Bias Stability	8	deg/h
	Angle Random Walk	0.12	deg/h
	Dynamic Range	± 100	deg/s
	−3 dB Bandwidth	350	Hz
Accelerometer	Nonlinearity	± 0.1	% of FS
	Dynamic Range	± 5	g
	−3 dB Bandwidth	350	Hz

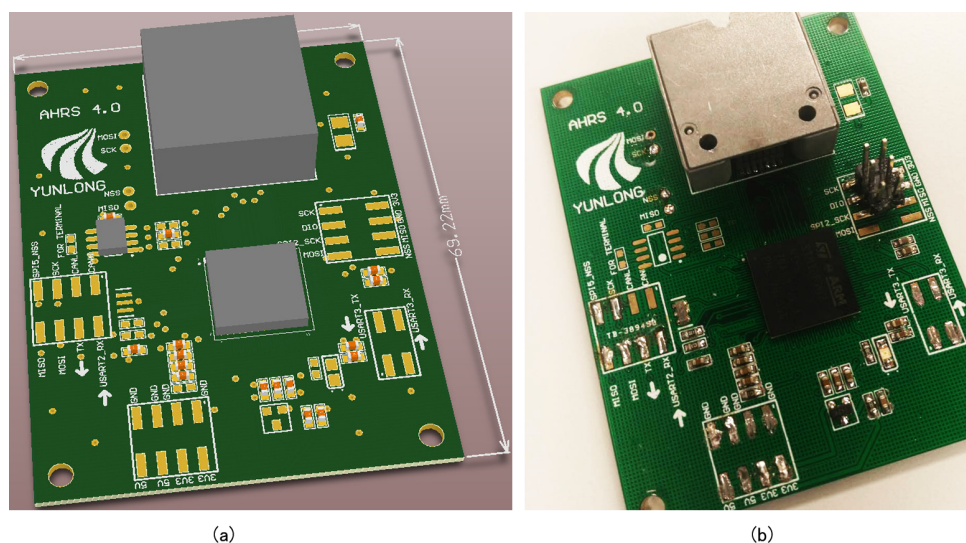


Figure 3. Attitude sensor board: (a) 3D view of the board in PCB design software; and (b) photo of the real board.

2.3. Stepper Motor Control and Drive Board

The stepper motor control and drive board is an important part of the attitude stabilization system, whose responsibilities are: (1) executing control algorithm with the frequency of 1 KHz; (2) amplifying the power of the signal from the controller to drive stepper motor; and (3) receiving and sending data to other boards or computers.

The stepper motor control and drive board is shown in Figure 4. There is one ARM chip (Cortex-M3 core) on this board with the CPU frequency of 168 MHz, which receives data from the attitude sensor board with UART port, runs the control algorithm and sends the controller outputs to stepper motor drive chips through SPI communication port. There are four layers on this board. The high voltage area on this board typically has a voltage of more than 20 V, while the low voltage area has a typical voltage of 3.3 V. These two areas are connected by a 0 ohm register, which can significantly decrease the electronic current affection from stepper motor. To avoid missing steps in stepper motor, the velocity profile is usually applied. There are many methods to implement the velocity profile by self-written code [21–23], but, in our case, a specific stepper motor control chip is applied. That is, the ARM chip sends the command rotation velocity to the stepper motor control chip, which is responsible to convert the command velocity into corresponding pulse, according to pre-defined velocity profile. The algorithms run in the attitude sensor board and stepper motor control and drive board are firstly designed in Matlab/Simulink, and then converted into C code by the code generation tools [24–26].

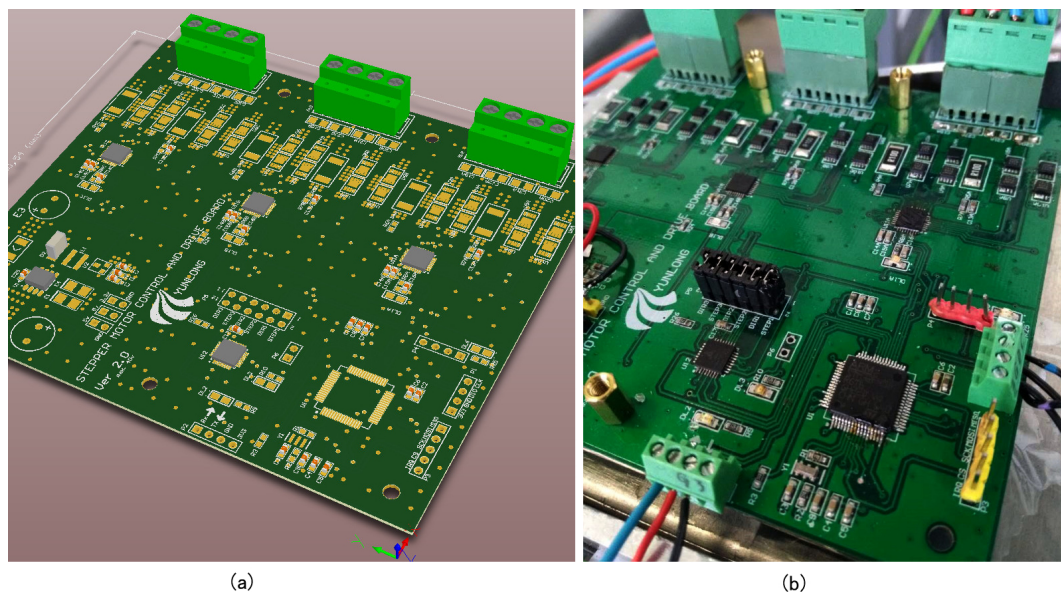


Figure 4. Stepper motor control and drive board: (a) 3D view of the board in PCB design software; and (b) photo of the real board.

3. Modeling and Parameter Identification

The first step in the controller design of attitude stabilization system of elevation part is obtaining the mathematical model of the elevation part. The velocity model is obtained first and the angle model can be obtained by the integration of the velocity model.

3.1. Velocity Model of Elevation Part

In the MSTA under development, the stepper motor is used as the actuator and is working in velocity mode. The velocity model of the elevation part is actually also the velocity model of the stepper motor. The reason is that the elevation part is considered as the load of the stepper motor.

The modeling method of the stepper motor proposed here is different from the conventional method. In the conventional method, the stepper motor works in open-loop mode and the rotation velocity of the shaft of stepper motor is unmeasurable. However, in our case, the rotation velocity of the shaft of stepper motor is measured by high-precision Micro-Electro-Mechanical Systems (MEMS) gyroscopes mounted together with the elevation part. The stepper motor model built in this part is simpler than that in previous papers, and can be easily used in simulation, linear and nonlinear controller design. Moreover, the parameters in the model can be easily identified from sampled experiment data while the parameters in the model in previous papers, such as the motor maximum magnetic flux, ψ_m , are very hard to be obtained.

The detailed procedure of modeling is stated next. In open loop mode, the controller sends out the command velocity, $V_c(t)$, to the stepper motor control and drive chips, which are responsible for rotating the magnetic field inside stepper motor. The rotation velocity of the magnetic field generated by the stator of stepper motor is denoted as $V_s(t)$, and the relationship between them is expressed in Equation (1),

$$\dot{V}_s(t) = \frac{1}{t_{mc}} (V_c(t) - V_s(t)) \quad (1)$$

where t_{mc} is a time constant of the first-order transfer function that is determined through experiments. If Equation (1) is transformed into frequency domain through Laplace transform, it becomes $\frac{V_s(s)}{V_c(s)} = \frac{1}{t_{mc} \cdot s + 1}$ and is actually a low-pass filter. The reason for using Equation (1) is that V_s cannot be changed sharply to avoid missing steps, which is fundamentally caused by the output torque limitation of the stepper motor. The change of V_s has to follow a velocity profile, which is achieved by the stepper motor control and driver chips.

The calculation of the output torque of stepper motor, $T_s(t)$ is illustrated in Figure 5.

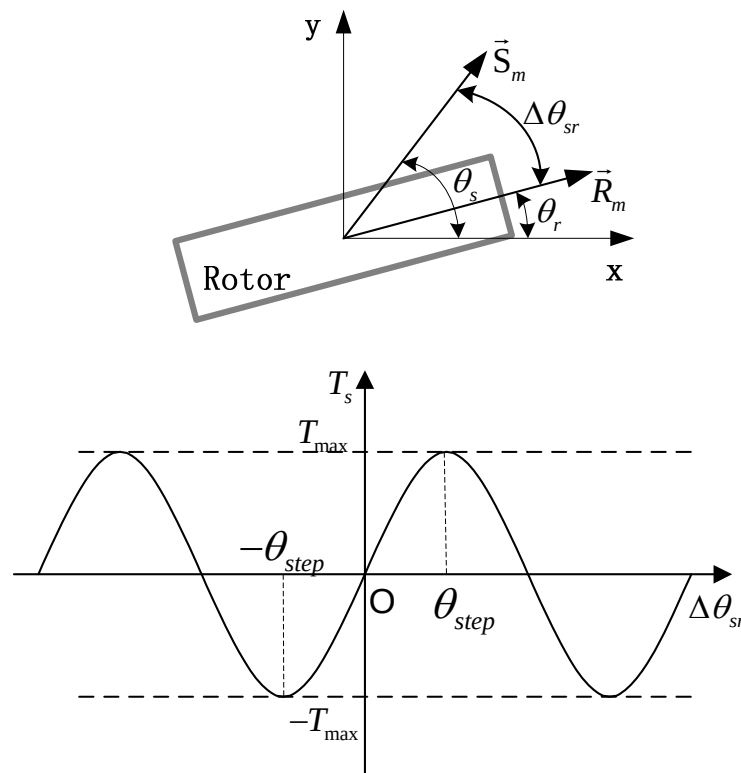


Figure 5. Calculation of the output torque of stepper motor.

In Figure 5, $\vec{S}_m$ is the direction of the magnetic field generated by the stepper motor stator, whose rotation angle is represented by θ_s . $\vec{R}_m$ is the direction of the magnetic field of the rotor, whose rotation angle is represented by θ_r . θ_{step} is the step angle and is $\frac{1.8 \cdot \pi}{180}$ rad. $T_s(t)$ is calculated by

$$T_s(t) = T_{\max} \cdot \sin \left(\frac{2\pi}{4 \cdot \theta_{step}} \cdot \Delta\theta_{sr}(t) \right) \quad (2)$$

$$\Delta\theta_{sr}(t) = \theta_s(t) - \theta_r(t) \quad (3)$$

$$\theta_s(t) = \int V_s(t) dt \quad (4)$$

$$\theta_r(t) = \int V_r(t) dt \quad (5)$$

where $T_{\max}$ is the maximum output torque of the stepper motor. Note that $T_{\max}$ is different from and is smaller than holding torque. In addition, note that $T_{\max}$ is not a constant but a curve of rotation velocity of motor shaft, as shown in Figure 6.

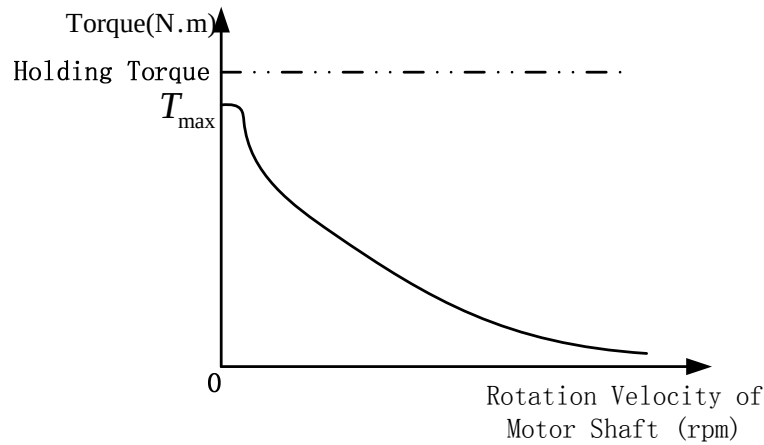


Figure 6. The curve of $T_{\max}$.

The curve of $T_{\max}$ in Figure 6 can be found in the stepper motor datasheet and is converted into a function of rotational velocity or data table for use in simulation or controller design. Because of the use of $T_{\max}$ curve, the back Electromotive Force (EMF) does not need to be considered. The reason is that $T_{\max}$ and back EMF have the same function. The amplitude of back EMF increases as the rotation velocity of motor shaft increases, which causes the decrease of the maximum phase current and further causes the decrease of the output torque of stepper motor. Using $T_{\max}$ is also better than using back EMF because $T_{\max}$ curve is provided by the stepper motor manufacturer. $T_{\max}$ is also the parameter that is used for checking whether the stepper motor has missed steps.

The dynamics of the mechanical system consisting of stepper motor and the load is expressed as,

$$T_s = J \cdot \dot{V}_r(t) + \zeta_m \cdot V_r(t) + T_L \quad (6)$$

where $J = J_M + J_L$ is the moment of inertia of the rotation part, and consists of moment of inertia of stepper motor rotor, J_m , and that of the load, J_L . J_L can be found from Computer-aided Design (CAD) design files or be obtained through experiment by using torque meter. Note that, when a belt transmission or gearbox is used, J_L needs to be divided by square of the speed reducing ratio for converting the moment of the inertia of the load into motor shaft. ζ_m is the damping ratio of the mechanical system and is obtained through experiments. T_L is the torque needed to hold the load when the system is in static condition, and is usually measured through a torque meter.

The nonlinear model of the stepper and load system is obtained by combining Equations (1)–(6).

3.2. Linear Stepper Motor Model

Currently, the controller design based on linear system has been well studied, whose stability can be easily proved and controller performance can be optimized by those linear optimization theory. Thus, linear controller is designed first to achieve the velocity control of the elevation part.

The first step of designing a linear controller is having a linear model of the plant.

In the nonlinear model of stepper motor, Equations (1)–(6), only Equation (2) is nonlinear and is linearized as below,

$$\begin{aligned} T_s &= T_{\max} \cdot \sin\left(\frac{2\pi}{4\theta_{\text{step}}} \cdot \Delta\theta_{sr}\right) \\ &\approx T_{\max} \cdot \frac{2\pi}{4\theta_{\text{step}}} \cdot \Delta\theta_{sr} \end{aligned} \quad (7)$$

$$= K_{\theta} \cdot \Delta\theta_{sr} \quad (8)$$

where $\sin(x) \approx x$ is applied. Furthermore,

$$K_{\theta} = \frac{T_{\max} \cdot \pi}{2\theta_{\text{step}}} \quad (9)$$

is the rotational stiffness.

Combining Equations (1), (3), (6) and (8), the linear velocity model of the motor and load system is obtained and is expressed as transfer function. The detailed procedure of this step is stated below.

T_L in Equation (6) is measured directly by the torque meter mounted together with the rotation shaft of the stepper motor, and is around 0.07 N·m. As T_L is so little, it is reasonable to remove it from Equation (6) to get the velocity transfer function, $G_v(s) = \frac{V_r(s)}{V_c(s)}$. Then, Equation (6) becomes

$$T_s = J \cdot \dot{V}_r(t) + \xi_m \cdot V_r(t) \quad (10)$$

By combining Equations (3)–(5), (8), and (10), we can get

$$\begin{aligned} K_{\theta} \cdot (\theta_s(t) - \theta_r(t)) &= J \cdot \dot{V}_r(t) + \xi_m \cdot V_r(t) \\ \Downarrow \\ K_{\theta} \left(\int V_s(t) dt - \int V_r(t) dt \right) &= J \dot{V}_r(t) + \xi_m V_r(t) \end{aligned} \quad (11)$$

By Laplace transform in Equation (11), we can get

$$\begin{aligned} K_{\theta} \cdot \left(\frac{V_s(s)}{s} - \frac{V_r(s)}{s} \right) &= J \cdot s \cdot V_r(s) + \xi_m \cdot s \cdot V_r(s) \\ \Downarrow \\ \frac{V_r(s)}{V_s(s)} &= \frac{w_n^2}{s^2 + 2 \cdot \xi_v \cdot w_n \cdot s + w_n^2} \end{aligned} \quad (12)$$

where

$$w_n = \sqrt{\frac{K_{\theta}}{J}}, \quad (13)$$

and

$$\xi_v = \frac{\xi_m}{2\sqrt{K_{\theta} \cdot J}} \quad (14)$$

$G_v(s)$ can be written as,

$$G_v(s) = \frac{V_r(s)}{V_c(s)} \quad (15)$$

$$= \frac{V_s(s)}{V_c(s)} \cdot \frac{V_r(s)}{V_s(s)} \quad (16)$$

$$= \frac{1}{t_{mc} \cdot s + 1} \cdot \frac{w_n^2}{s^2 + 2 \cdot \zeta_v \cdot w_n \cdot s + w_n^2} \quad (17)$$

3.3. Parameters Identification

In previous section, the linear model of stepper motor and load system are obtained, and, in this section, the parameters of these models are estimated through experiments or through calculation. The parameters that can be obtained through experiments are t_{mc} , ζ_v , and w_n . The parameters that can be obtained through calculation are K_θ , ζ_m , and w_n . Note that w_n , an important parameter in the model, can be obtained both ways.

3.3.1. t_{mc}

The most significant difference between the stepper motor model in this paper and those in previous papers lies in Equation (1), which allows us to ignore those hard parameters inside the stepper motor, such as phase current, back EMF and magnetic flux. Only one parameter, t_{mc} , needs to be considered. The rotation velocity of the magnetic field of stator of stepper motor cannot be measured directly and is obtained indirectly through measurements of phase current of stepper motor. The command rotation velocity can be easily obtained by measuring the pulse input to the stepper motor. By comparing these two velocities, t_{mc} is estimated to be about 1 ms.

3.3.2. ζ_v and w_n

In this section, w_n is firstly calculated and then ζ_v and w_n are estimated using the experimental data. From Equations (9) and (13), we can get,

$$w_n = \sqrt{\frac{K_\theta}{J}} = \sqrt{\frac{T_{max} \cdot \pi}{2\theta_{step} \cdot J}}$$

where $T_{max} = 1.85 \text{ N}\cdot\text{m}$, $\theta_{step} = \frac{1.8\pi}{180} \text{ rad}$, and $J = 0.08 \text{ kg}\cdot\text{m}^2$. Thus, $w_n \approx 30.4 \text{ rad/s}$.

$t_{mc} = 0.001 \text{ s}$ is obtained in the last part, which means the -3 dB bandwidth of $\frac{1}{t_{mc} \cdot s + 1}$ is about $\frac{1}{t_{mc}} \approx 1000 \text{ rad/s}$ and is more than 10 times larger than w_n . This means $\frac{1}{t_{mc} \cdot s + 1}$ can be removed from Equation (15), and a simpler version of $G_v(s)$, $G_{vs}(s)$, is expressed as,

$$G_{vs}(s) = \frac{w_n^2}{s^2 + 2 \cdot \zeta_v \cdot w_n \cdot s + w_n^2}, \quad (18)$$

which is the final model used for model identification.

Many experiments were done to obtain the data for model identification. The command velocity is 35 deg/s , which, together with the measured velocity, is shown in Figure 7.

From the experimental data, ζ_v and w_n can be obtained through one of those numerical optimization methods, such as gradient descent methods or Gauss–Newton methods. The identification results are $\zeta_v = 0.022$ and $w_n = 27.3 \text{ rad/s}$. The output of the model in Equation (18) is compared with the experimental data, as shown in Figure 7, which verifies the correctness of the model.

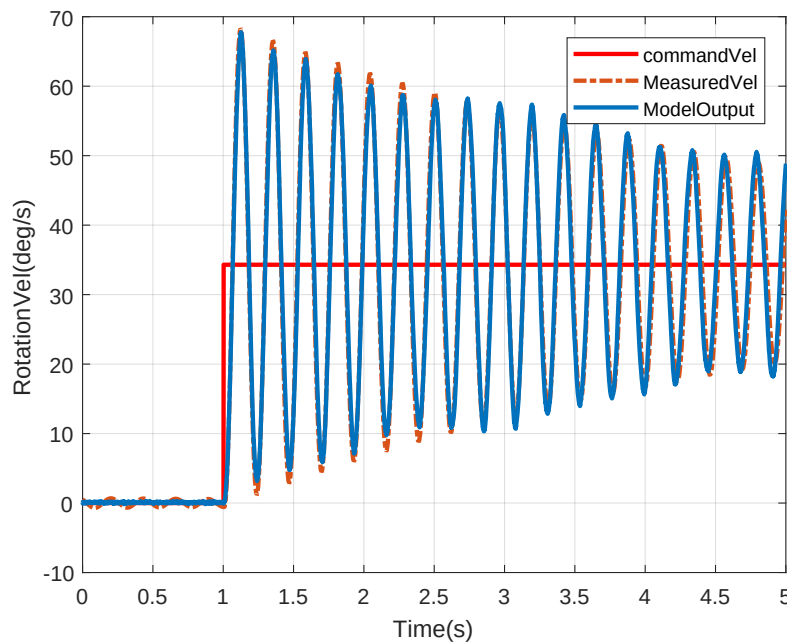


Figure 7. Experimental data and model output.

3.4. Model Uncertainty

Although the error between model output and experiment data is little in Figure 7, it should be noted that the error could be obvious at different command velocities. As shown in Figure 8, when the command velocity is changed to 3.5 deg/s, the error between the output of the model identified in Section 3.3 and the experiment data is obvious. This means the problem of model uncertainty cannot be ignored, which is the main reason for applying $\mathcal{H}_\infty$ loop shaping method, a kind of robust control, for the velocity controller design. In addition, note that a nonlinear model can represent the real plant with high accuracy, but our design techniques cannot effectively deal with it [27]. The model in Section 3.3 is used as the nominal model in robust controller design.

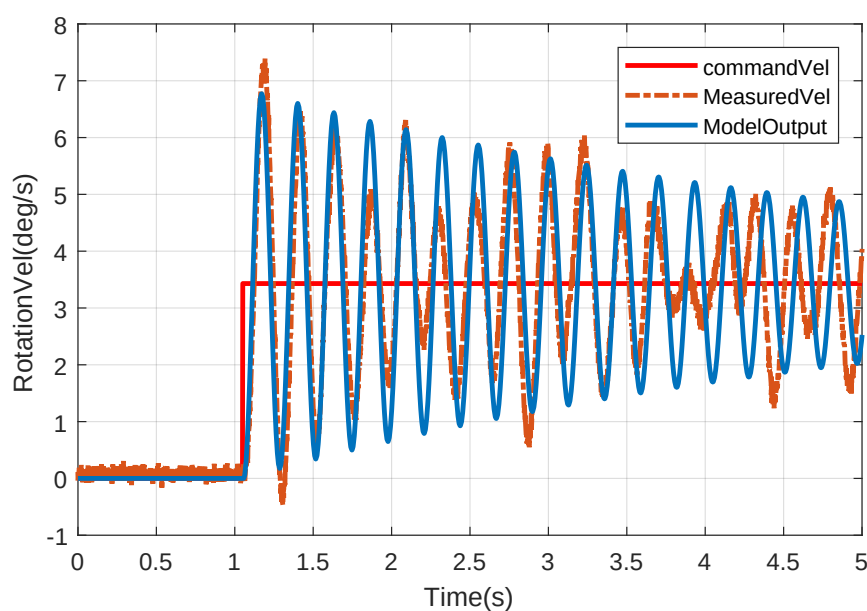


Figure 8. Model uncertainty in different command velocity.

4. $\mathcal{H}_\infty$ Loop-Shaping Design

$\mathcal{H}_\infty$ loop shaping is a popular controller design method that has been successfully applied in many fields [28–31]. The loop-shaping design procedure described in this section is based on H_∞ robust stabilization combined with classical loop shaping, which was proposed by McFarlane and Glover [32–34]. It consists of two steps. First, the open-loop plant is augmented by pre- and post-compensators to give a desired shape to the Bode magnitude plot of the open-loop frequency response. Then, the resulting shaped plant is robustly stabilized with respect to coprime factor uncertainty using $\mathcal{H}_\infty$ optimization [35]. Note that this coprime factor uncertainty does not use the uncertainty information from the plant. The coprime uncertainty description provides a good “generic” uncertainty description for cases where we do not use any specific a priori uncertainty information [35].

The plant, expressed in Equation (18), is pre- and post-compensated, respectively, by W_1 and W_2 , as shown below,

$$G_s = W_1 \cdot G_{vs} \cdot W_2 \quad (19)$$

where W_1 is the pre-compensator, W_2 is the post-compensator, and G_s is the shaped plant.

The main function of attitude stabilization system is for disturbance rejection. According to the product performance requirements, the RMS value of pitch tracking error should be less than 0.2 deg while the disturbance from ship is,

$$D_s = A \cdot \sin\left(\frac{2\pi}{T} \cdot t\right) \quad (20)$$

where $A = 25$ deg and $T = 6$ s.

In our case, W_1 was set to be the inverse of G_{vs} to counteract the high open-loop gain of G_{vs} at w_n . Then, W_2 was selected to get acceptable disturbance rejection while keep the system stable. To get good disturbance rejection, the loop gain, $|G_s(j\omega)|$, should be as large as possible within the bandwidth region. For the existence of unmodeled high-frequency dynamics and limitations on the allowed manipulated inputs, $|G_s(j\omega)|$ has to drop below one at and above the crossover frequency ω_c . After trial and error, W_2 was set to be,

$$W_2 = \frac{6.8 \cdot (s^2 + 13.3s + 361)}{s^2 \cdot (s^2 + 45s + 2025)} \quad (21)$$

where the phase lead compensator, $\frac{s^2+13.3s+361}{s^2+45s+2025}$, are applied for enhancing the responsiveness and stability of the system.

The Bode plot of the frequency response of G_s is shown in Figure 9.

Next, the procedure of $\mathcal{H}_\infty$ robust stabilization of G_s is stated.

The perturbed plant model of the nominal plant, G_s , is expressed as,

$$G_{sp} = (M_l + \Delta_{M_l})^{-1}(N_l + \Delta_{N_l}) \quad (22)$$

where M_l and N_l are stable coprime transfer functions obtained from left coprime factorization of G_s , that is, $G_s = M_l^{-1} \cdot N_l$. Δ_{M_l} and Δ_{N_l} are stable unknown transfer functions which represent the uncertainty in G_s .

The family of perturbed plants that should be stabilized are defined as,

$$G_{sp} = \left\{ (M_l + \Delta_{M_l})^{-1}(N_l + \Delta_{N_l}) : \left\| \begin{bmatrix} \Delta_{N_l} & \Delta_{M_l} \end{bmatrix}_\infty < \epsilon \right\| \right\} \quad (23)$$

where $\epsilon > 0$ is the stability margin. The objective of robust stabilization is to maximize ϵ .

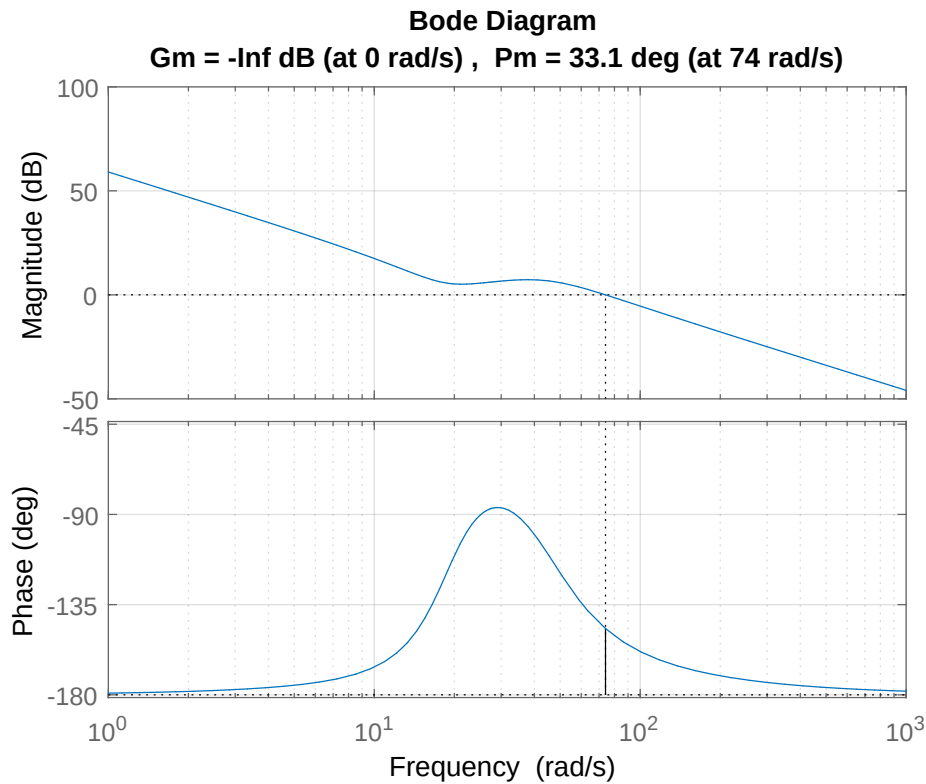


Figure 9. Bode plot of the frequency response of G_s .

The perturbed feedback system of velocity loop is shown in Figure 10, where u is the manipulated variable, y is the rotation velocity of elevation part and is also the measurement of gyroscope, and K_{sv} is the velocity controller for shaped plant.

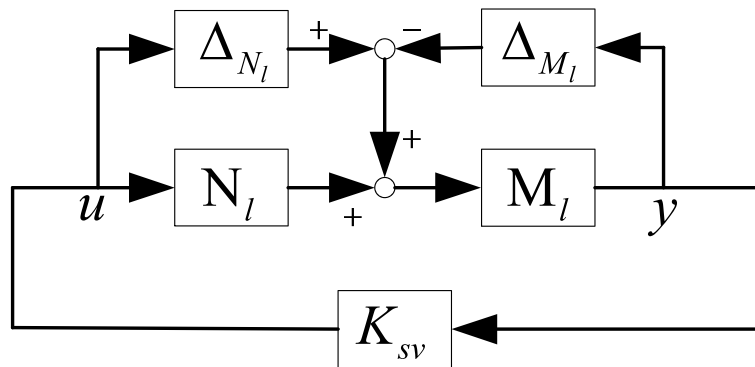


Figure 10. $\mathcal{H}_\infty$ robust stabilization of velocity loop.

The stability property is robust if and only if the nominal feedback system is stable and

$$\gamma \triangleq \left\| \begin{bmatrix} K_{sv} \\ I \end{bmatrix} (1 - G_s K_{sv})^{-1} M_l^{-1} \right\|_\infty \leq \frac{1}{\epsilon} \quad (24)$$

Getting the maximum ϵ also means getting the minimum γ , denoted as $\gamma_{\min}$, which is obtained by

$$\gamma_{\min} = \epsilon_{\max}^{-1} = (1 + \rho(XZ))^{\frac{1}{2}} \quad (25)$$

where ρ is the maximum eigenvalue, and for a minimal state-space realization (A, B, C) of G_{sp} , and X and Z are the unique positive definite solutions of the algebraic Riccati equations, Equations (26) and (27), respectively. Note that $\gamma_{\min}$ is the inverse of the magnitude of coprime uncertainty we can tolerate before we get instability.

$$A^T X + XA - XBS^{-1}B^T X + C^T R^{-1}C = 0, \quad (26)$$

Z is the unique positive definite solution of the following algebraic Riccati equation,

$$AZ + ZA^T - ZC^T CZ + BB^T = 0. \quad (27)$$

Then, the K_{sv} that guarantees that

$$\left\| \begin{bmatrix} K_{sv} \\ I \end{bmatrix} (1 - G_s K_{sv})^{-1} M_l^{-1} \right\|_{\infty} \leq \gamma, \quad (28)$$

where $\gamma = 1.1$, $\gamma_{\min}$ is typically used, is given by

$$A_{K_{sv}} = A - BB^T X + \gamma^2 (L^T)^{-1} Z C^T C \quad (29)$$

$$B_{K_{sv}} = B^T X \quad (30)$$

$$C_{K_{sv}} = \gamma^2 (L^T)^{-1} Z C^T \quad (31)$$

where $L = (1 - \gamma^2)I + XZ$, and $(A_{K_{sv}}, B_{K_{sv}}, C_{K_{sv}})$ is the minimum state-space realization of K_{sv} . The velocity controller, K_v , for the plant, G_{vs} , is $K_v = W_1 K_{sv} W_2$.

In our case,

$$A = \begin{bmatrix} -46.20 & -44.13 & -17.56 & -23.03 & 0 & 0 \\ 64 & 0 & 0 & 0 & 0 & 0 \\ 0 & 32 & 0 & 0 & 0 & 0 \\ 0 & 0 & 32 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0.06 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0.03 & 0 \end{bmatrix}, \quad (32)$$

$$B = \begin{bmatrix} 4096 & 0 & 0 & 0 & 0 & 0 \end{bmatrix}^T, \quad (33)$$

$$C = \begin{bmatrix} 0 & 0.02 & 0.01 & 0.02 & 3.11 & 2590.73 \end{bmatrix}. \quad (34)$$

The specific expression of K_{sv} and K_v , respectively, are

$$K_{sv} = \frac{511.37(s + 22.09)(s^2 + 32.74s + 982.6)(s^2 + 1.201s + 745.3)}{(s^2 + 13.29s + 371.6)(s^2 + 1.201s + 745.3)(s^2 + 429.6s + 53780)}, \quad (35)$$

$$K_v = \frac{3463.9(s + 22.09)(s^2 + 13.3s + 361)(s^2 + 32.74s + 982.6)(s^2 + 1.201s + 745.3)}{s^2(s^2 + 13.29s + 371.6)(s^2 + 45s + 2025)(s^2 + 429.6s + 53780)}. \quad (36)$$

and $\gamma_{\min}$ is 1.89, corresponding to 52.91% allowed coprime uncertainty.

The Bode plot of the velocity open-loop frequency response after robust stabilization is shown in Figure 11. Compared with Figure 9, it can be seen that the phase margin is enlarged in Figure 11 for more robust stabilization, and the system bandwidth is decreased from 74 rad/s to 49.1 rad/s, which shows the conservative characteristic of robust control. That is, the control performance will be decreased for the increase of system robustness.

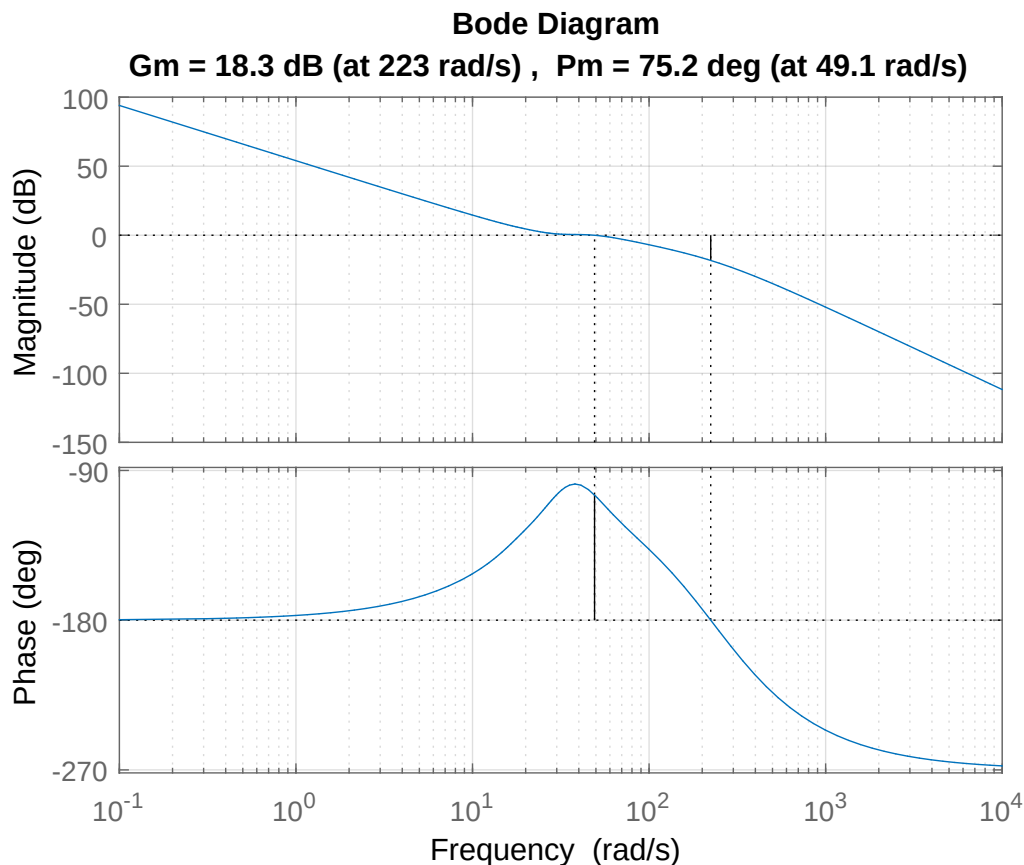


Figure 11. Bode plot of the frequency response of $K_v G_{vs}$.

5. Experiment Setup and Results

The controller designed in Section 4 was tested on an experimental system, as shown in Figure 12. A Stewart platform was used as a ship simulator for disturbance generation. The rotation movement of the Stewart platform is the sinusoidal function in Equation (20). The elevation part of the MSTA was installed on a test bench, where the rotation torque, motor current, rotation velocity and angle can be measured. The robust velocity controller together with the PI position controller were converted into discrete form and implemented in C code, which is put in the Interrupt Service Routine (ISR) of the timer interrupt in ARM chip. The timer interruption period was set to be 1 ms. Several important variables, such as current rotation angle, rotation velocity, reference angle, and torque, were sent to the computer through UART port and displayed in Simulink in real-time mode. During the experiment, the reference pitch was 50 deg and the running time was about 85 s. The running results are shown in Figure 13.

In Figure 13, it can be seen that the designed robust controller can stabilize the antenna dish with high precision, which makes the two curves shown in the top part of Figure 13 almost overlapped. The angle error between the reference pitch and measured pitch is shown in the bottom part of Figure 13 and the pitch error is in the range of $[-0.15 \ 0.15]$ deg, whose RMS value is definitely less than the 0.2 deg performance requirement of Ka-band MSTA. The rotation velocity of antenna dish is shown in Figure 14 to give more information about the running states of the system. From the rotation velocity curve, it can be seen that, most of the time, the rotation velocity of antenna dish is in the range of $[-2 \ 2]$ deg/s while the maximum velocity disturbance from Stewart platform can reach about 26.2 deg/s, which again shows the high performance of designed control system.



Figure 12. Experimental system for testing robust controller.

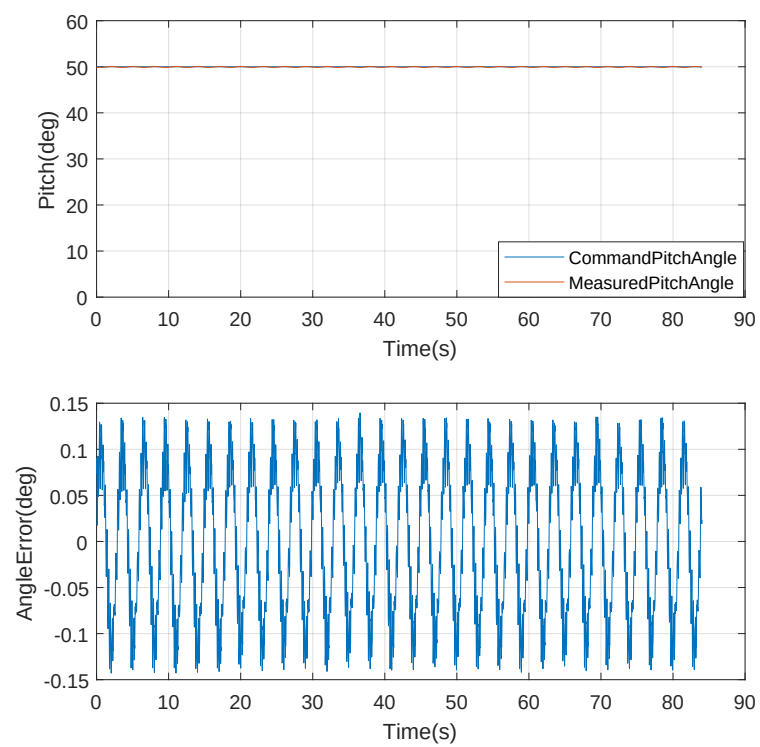


Figure 13. Running results of attitude stabilization system of MST.

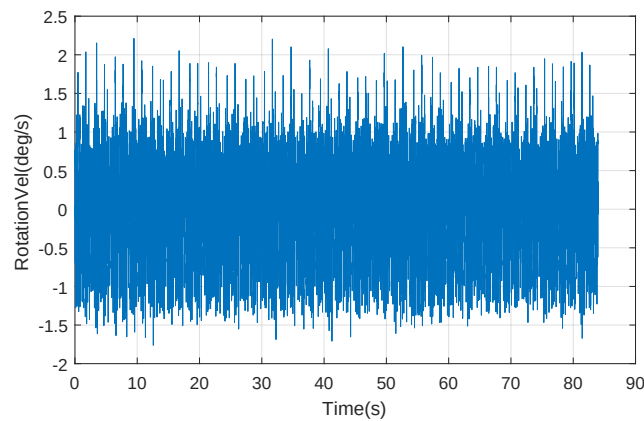


Figure 14. Rotation velocity of antenna dish during experiment.

Note that the controller designed above is actually only one of the two controllers for this system. In the current stage, when the MSTA is started, a normal Proportional-Integral-Derivative (PID) controller is firstly applied to adjust the antenna dish to the command attitude, and then the controller in this paper is switched on, which means the angle error is actually very small when the controller in this paper starts working. In addition, note that the data shown in Figures 13 and 14 are sampled when the system has already reached the stable state, which is why we cannot see the transient stage at the beginning of these figures.

In this experiment, only sinusoid input was applied, the frequency (1 rad/s) and amplitude (25 deg) of which correspond to the worst case specified in the product performance requirements. In this worst case, the maximum ship rotation velocity was 26.2 deg/s and the control system could keep the RMS value of tracking angle error less than 0.2 deg. According to loop gain in the bode diagram in Figure 11, the designed robust controller could reject any combination of irregular input if the rotation velocity of the input was less than 26.2 deg/s, which was validated through the simulation experiment stated below in detail.

In this simulation, the disturbance input of pitch was set as,

$$Input = A_1 \sin\left(\frac{2\pi}{T_1} \cdot t\right) + A_2 \sin\left(\frac{2\pi}{T_2} \cdot t\right) + A_3 \sin\left(\frac{2\pi}{T_3} \cdot t\right)$$

where $A_1 = 5, A_2 = 5, A_3 = 5$ (unit: deg/s) and $T_1 = 10, T_2 = 8, T_3 = 6$ (unit: s). These values were chosen with the requirement that the rotation velocity of input should not be larger than 26.2 deg/s. The curve of the input is shown in Figure 15.

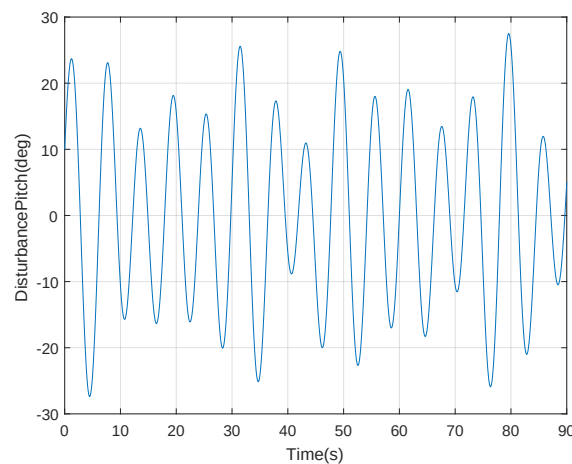


Figure 15. Disturbance input of pitch.

With this input, the controller performance can be seen in Figure 16. In the upper part of Figure 16, the angle error between command (50 deg) and the measurement is shown. In this figure, it can be seen that, at the beginning, the angle error is relatively large, but the error quickly converges and reaches a stable state after around 10 s. The angle error in the stable state (after 20 s) is shown in the bottom part of Figure 16 to show more details, from which it can be seen that the angle error is less than 0.1 deg.

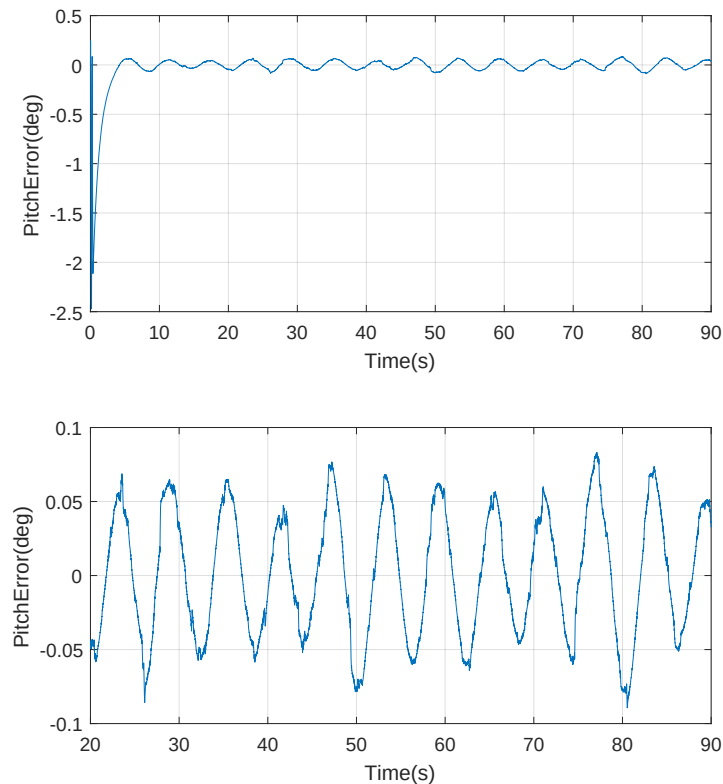


Figure 16. Running results of attitude stabilization system of MSTA in simulation environment.

6. Conclusions

A high performance attitude stabilization system is designed and implemented in this paper for the Ka-band MSTA. The system diagram is given and two important self-made electronic boards, the attitude sensor board and stepper motor control and drive board, are introduced with details. The robust velocity controller based on $\mathcal{H}_\infty$ loop shaping is designed and the detailed design procedure is given. The designed attitude stabilization system was tested on a Stewart platform, which was used as a ship simulator. From the experimental results, it can be seen that the performance of designed attitude system is very high and the tracking angle error is less than 0.15 deg when the maximum velocity disturbance from Stewart platform is 26.2 deg/s. The performance of the designed robust controller has only been validated in the laboratory and not in the real ship environment, where harsher conditions can be expected. Thus, the designed controller needs to be further tested on a real ship to further validate its performance. Moreover, the MSTA usually continuously works for several months. In such case, the properties of the electronic and mechanical components would change, which can result in the variation of the nominal model used for robust controller design. Thus, a system performance decrease can be expected, which has not been analysed yet. Furthermore, only the pitch control loop is designed and implemented in this paper. However, there are actually totally three control loops (roll, pitch and yaw). The control loop design and implementation of the roll and yaw, together with the coupling effects between them, are left for future works.

Author Contributions: Y.W. contributed to defining the overall problem and proposed the core scientific idea to solve it. Y.W. student designed and implemented the hardware experiments to validate the proposed strategy. Y.W. wrote the entire draft version of the paper, and revised it according to co-authors' comments. M.S. assisted in careful reviewing of the paper and proposed various refinements to the draft proposal made by the PhD student. D.M.A.H. helped review and improve the paper. R.M.C. had many valuable discussions with Y.W. about the control system design.

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