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A NARROW-BAND TRACKING FILTER

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Supplement to TN D-1419

A NARROW-BAND TRACKING FILTER

by

Ronald Woodman*
Network Engineering Branch
Network Engineering and Operations Division

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Supplement to TN D-1419, "A Phase-Locked Phase Filter for the Minitrack System"

*Mr. Woodman is employed by the Instituto Geofisico Del Peru, Lima, Peru

ABSTRACT

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The characteristics of a servo phase-lock tracking filter is described herein. The filter tracks a 100 cps carrier with a bandwidth of either 0.3 cps or 0.03 cps. The advantages of this type filter and its use is also discussed.

Author

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This report is to supplement the NASA TN D-1419 report, "A Phase-Locked Phase Filter for the Minitrack System", to cover the improved prototype built by the Network Engineering Branch, Data Systems Section.

After due consideration, the name "Phase-Locked Phase Filter" has been changed to "Narrow Band Tracking Filter". This name implies its function and the technique it uses.

The block diagram and system theory is essentially the same as described in the above Technical Note. There were two different approaches that could have been taken: to shift the reference and phase-lock it to the signal or to shift the signal and phase-lock it to the reference. The first approach has been taken since the output is a clean 100 cps signal whose phase can be measured and digitalized with existing Minitrack equipment.

The system has been designed for two different bandwidths: 0.3 cps with 10 sec. acquisition time and 0.03 cps with 100 sec. acquisition time. The dynamics of the system are slightly different than the dynamics of the system described in the above mentioned TN report. The open loop transfer function is still of the form.

$$G(s) = \frac{K(s + z)}{s^2(s + p)} \tag{1}$$

but the ratio p/z has been increased from 5 to 9. A ratio of nine generates a root locus as shown in Figure 1, where for a particular gain (K = p/3) the three poles of the close loop transfer function lie on the same point of the real axis. This results in a critically damped second order servo with 40 db/sec roll-off. A value for z equal to 0.371 or 0.0371 has been chosen to give a bandwidth of 0.3 or 0.03 cps and a acquisition time of 10 or 100 seconds. The resulting closed loop transfer function is

$$\frac{\phi_u(s)}{\phi_i(s)} = \frac{0.261\ (S + 0.371)}{(S + 1.125)^3} \qquad \text{for 0.3 cps BW} \tag{2}$$

and

$$\frac{\phi_o(s)}{\phi_1(s)} = \frac{0.00261\ (S + 0.0371)}{(S + 0.1125)^3} \quad \text{for } 0.03 \text{ cps BW} \tag{3}$$

with a frequency response as plotted in Figure 2 and a time response as shown in Figure 3. The frequency response shows a peak of 3 db as compared with 4 db with the former system and a step input response with 25% overshoot as compared to 30% (Figure 3).

It has been mentioned in the report that the steady state error for this filter is zero for a step as well as for velocity input. If the input has some acceleration there is some error that should be corrected at the time of processing the information. In most cases this acceleration is so small that it can be neglected. Nevertheless for an input of the form

$$\phi_i = a + bt + ct^2 \tag{4}$$

the output would be in error by

$$\phi_i - \phi_o = \frac{2C}{K_a} \tag{5}$$

where

$$K_a = s^2 G(s) \Big|_{s=0} = \frac{3.72(s + 0.371)}{(s + 3.34)} \Big|_{s=0} = 0.414$$

for 0.3 cps, and

$$K_a = 0.0372\ \frac{(s + 0.0371)}{(s + 0.334)} \Big|_{s=0} = 0.00414$$

for 0.03 cps. The error constant is proportional to the square of the bandwidth; fortunately the acceleration is proportional to the square of the maximum phase rates. If the 0.3 cps is limited to phase rates of the order of 0.5 lobes/sec (0.05 lobes/sec for the 0.03 cps), corrections, if any, of only a few counts are necessary. (One count equals .001 of 360° electrical degrees or .36°).

The phase shifter in the new design has been improved for optimum linearity. The outputs of both rotor windings are being summed exactly 90° out of phase and equal amplitude by a high gain operational amplifier with unity feedback. It was noticed that the linearity of the phase shifter was affected by any slight distortion of the signal. Distortion has been kept as low as possible and the signal is being filtered from harmonics before phase detection by a 15 cps band pass filter. A measured linearity of 0.05% has been achieved.

The AC operational amplifier was especially designed for high gain (90 db) at 100 cps and stable operation with feedback for unity gain. The phase shifter design is being used for other applications and will be described in a separate report.

A limiter stage has been introduced in the signal path before phase detection to get rid of amplitude variations. This limiting produces some noise quieting and generates a square wave for the phase detector.

The phase detector is the diode switching type. The signal square wave switches the 100 cps reference. The output of the detector after being filtered by the first operational amplifier is essentially the cross-correction of both inputs. The function of the DC Operational amplifiers are essentially the same. The values of some of the elements have been changed to conform with the new transfer function.

The Chopper, Servo Amplifier and Motor-Generator is a Diehl integrated package. The Motor-Generator is a 60 cps, 3,000 rpm maximum speed type with 7.5 in-oz stalled torque. A 10 in-oz^2 flywheel was clamped on the motor shaft to increase its mechanical time constant. A larger mechanical to electrical time constant ratio allows stronger feedback in the velocity loop, thereby reducing the effects of mechanical friction and increasing its dynamic range. With the flywheel, it was found that the amount of feedback was not limited by the loop stability but by amplifier saturation with generator residual noise. Therefore, the flywheel inertia effect is not being fully realized. A flywheel inertia of 3 in-oz^2 should be sufficient.

A problem has been encountered which was not noticed before. For a particular phase input signal the system oscillates very slowly, as much as ±2.5 count. This occurs only in the 0.03 cps. The oscillations occur for certain values of the fixed phase input which correspond to a particular position of the motor-generator rotor. If the system is tracking at a normal slow rate the problem does not develope since the rotor is never at the critical position long enough for the oscillations to build up. The only time, in practical use, that a fixed phase is fed into the system is during calibration. It is possible to adjust the phase shifter so that the calibration phase corresponds to a stable motor-generator rotor position, or calibrate in the 0.3 cps BW to avoid the oscillations.

The cause of the problem is the in-phase, position dependent, fundamental residual voltage of the motor-generator. The output of an ideal generator is zero for zero rpm. Actual generators have an output at 0 rpm which is called "residual voltage". The residual voltage is the sum of random noise and signals of the fundamental frequency and harmonics. The fundamental residual voltage has an "in-phase" and a "quadrature" component. It is only the in-phase component that causes any dynamic effects in the velocity servo loop. The in-phase residual voltage has a fixed and a position dependent component. The only effect of the fixed component is to cause "creeping" of the motor-generator, and in a type 2 position servo loop is automatically biased out. The in-phase, position dependent, fundamental residual component varies sinusoidally as a function of the position of the rotor, with as many cycles as number of poles in the generator. It is this last component that causes the oscillations, being position dependent, and when fed back behaves as a position feedback either degenerative or regenerative, depending on what part of the sinusoid the rotor is positioned. In the 0.03 cps mode, this regenerative feed is of the same order as the desireable position feedback of the loop causing the system to be unstable and oscillate. The magnitude of the oscillations are as large as the negative slope portion of the sinusoid, i.e., one fourth of a turn for a two pole generator. In this system there is a 50 to 1 gear reduction from the motor-generator to the resolver, therefore, a one fourth turn peak to peak oscillation corresponds to 0.005 turns peak to peak oscillation at the resolver or ±2.5 counts. If a 100:1 reduction were used the oscillations would be reduced to an acceptable ±1.25 counts with a reduction in maximum tracking rate from 0.5 lobes/sec to 0.25 lobes/sec which still would take care of satellite passes with normal phase rates. If 400 cps were used, the oscillations would be reduced by a factor of 0.15 (= 60/400) for corresponding tracking rates.

In order to get rid of the oscillations completely, it has been calculated that a motor-generator with a maximum in-phase, position dependent, residual voltage, e_m, less than:

$$e_m \quad \pm \left(0.003 \frac{v}{\pi p}\right) \quad \text{volts}$$

is necessary, where, v = output volts per 1000 rpm

p = number of poles

For a two pole unit and 6 v/1000 rpm

$$e_m = \frac{1}{2} \left(0.003 \frac{v}{\pi p}\right) = \frac{0.003 \times 6}{2 \times 3.14 \times 2} = 0.0014 \text{ volts}$$

would be necessary. Diehl Manufacturers state that they are in the development stage of a 60 cps unit that would meet such specifications.

Another solution for the problem is the use of a DC tachometer which will be tried in the laboratory. Increased friction and ripple on the output are new problems that would have to be solved.

The simplest solution and the one recommended for future units is to increase the gear reduction from 50 to 100 to 1.

A synchronous amplitude detector has been installed in the system with good results. The amplitude of the 100 cps signal input is synchronously detected using the output of the filter shifted 90° as the reference. The advantage of a synchronous detector over an envelope detector is that no noise is introduced in the process of detection and its threshold is determined by the post-detection bandwidth rather than the pre-detection bandwidth. Post-detection bandwidth can be set to either 0.3 or 0.03 cps.

Another advantage of the detector is that it detects the amplitude of the 100 cps output of the Minitrack receivers. This amplitude depends only on the magnitude of any coherent signal or noise going into both front ends of the receiver and discriminates against any noncoherent noise such as the one generated in each front end or the one coming from the sky background.

The detected amplitude could be used to control the AGC of the receiver. A well balanced product detector would not produce any 100 cps ripple which would be a problem when a fast AGC is desireable.

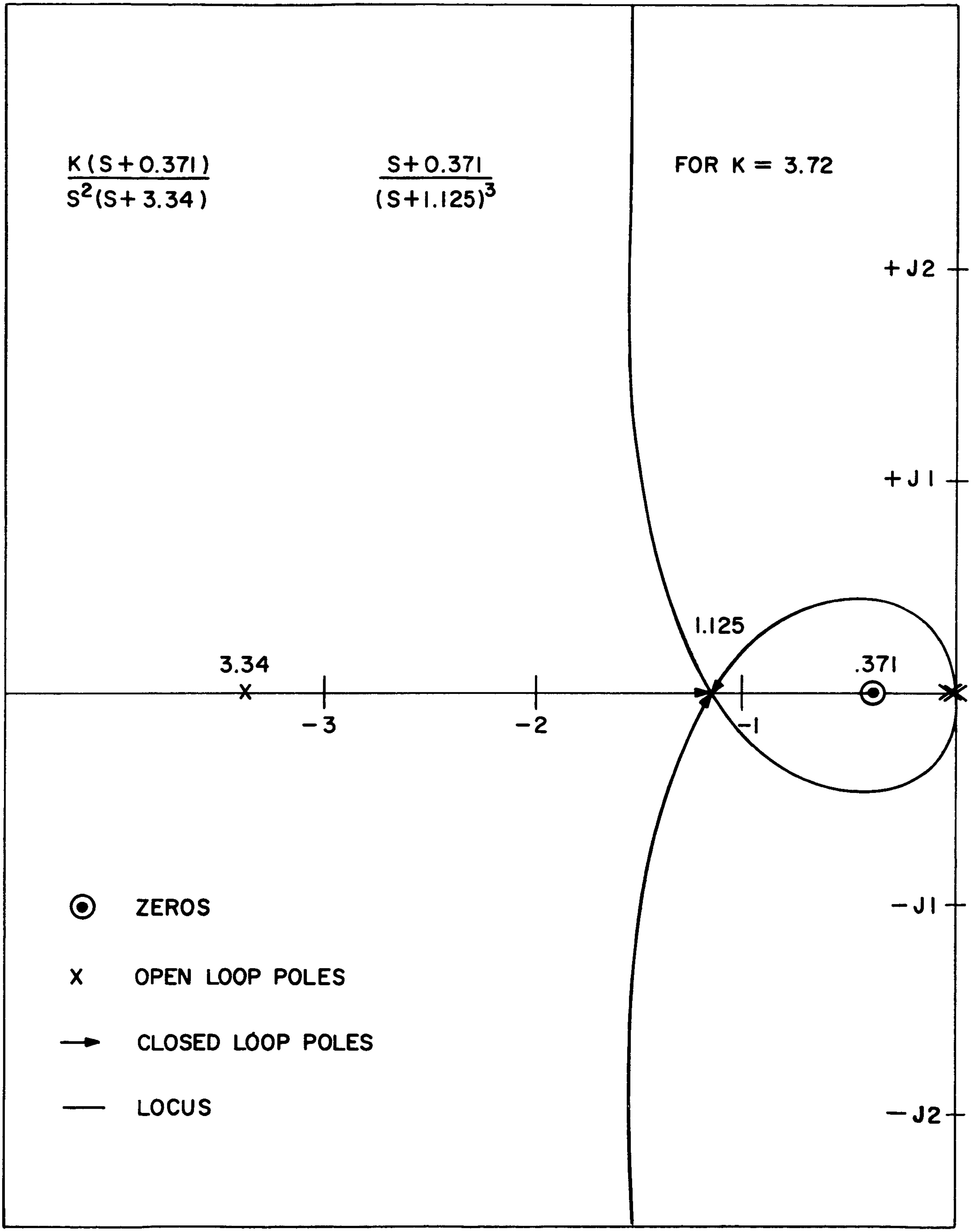


FIG.1 OPEN AND CLOSED LOOP ROOT LOCUS

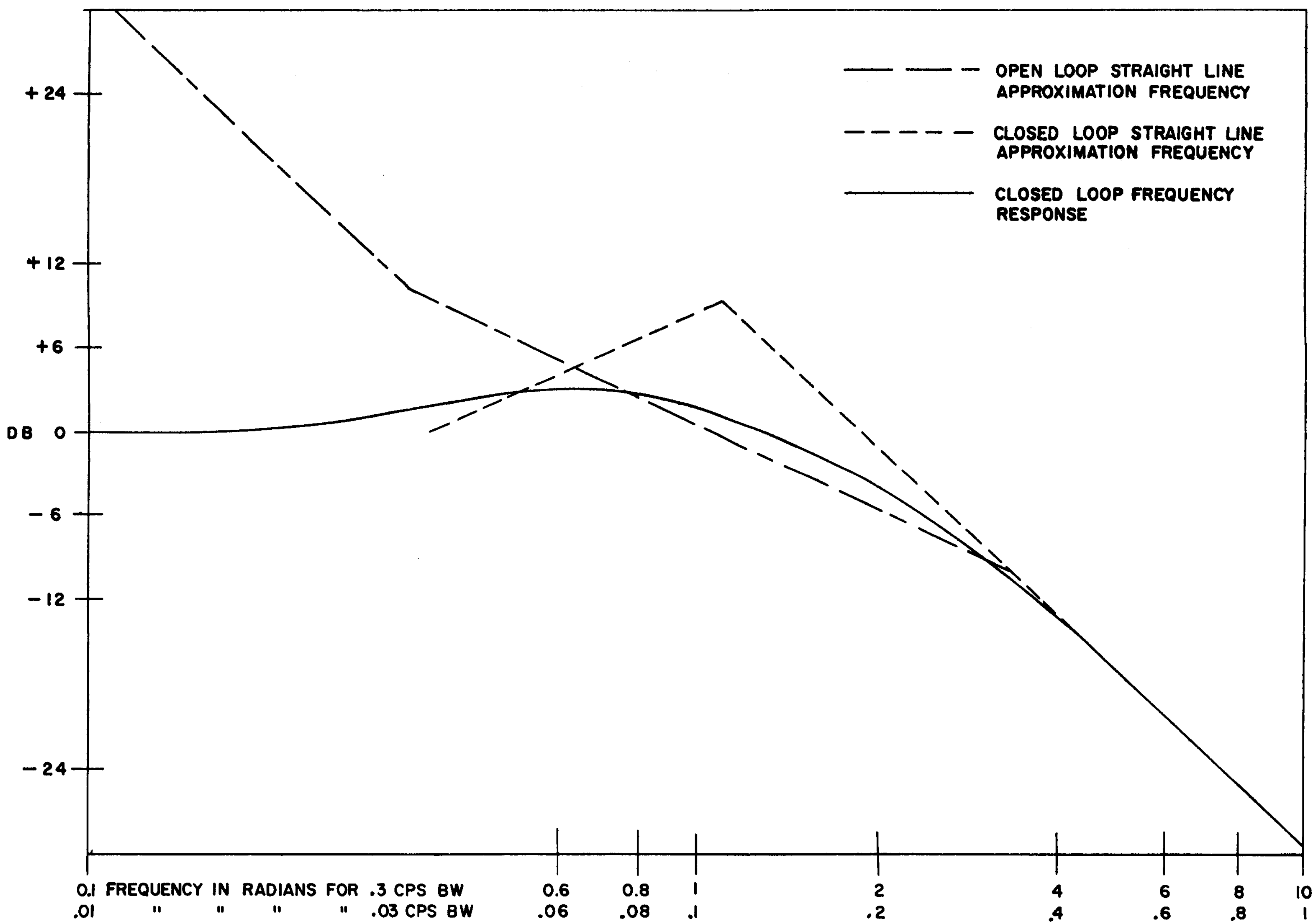
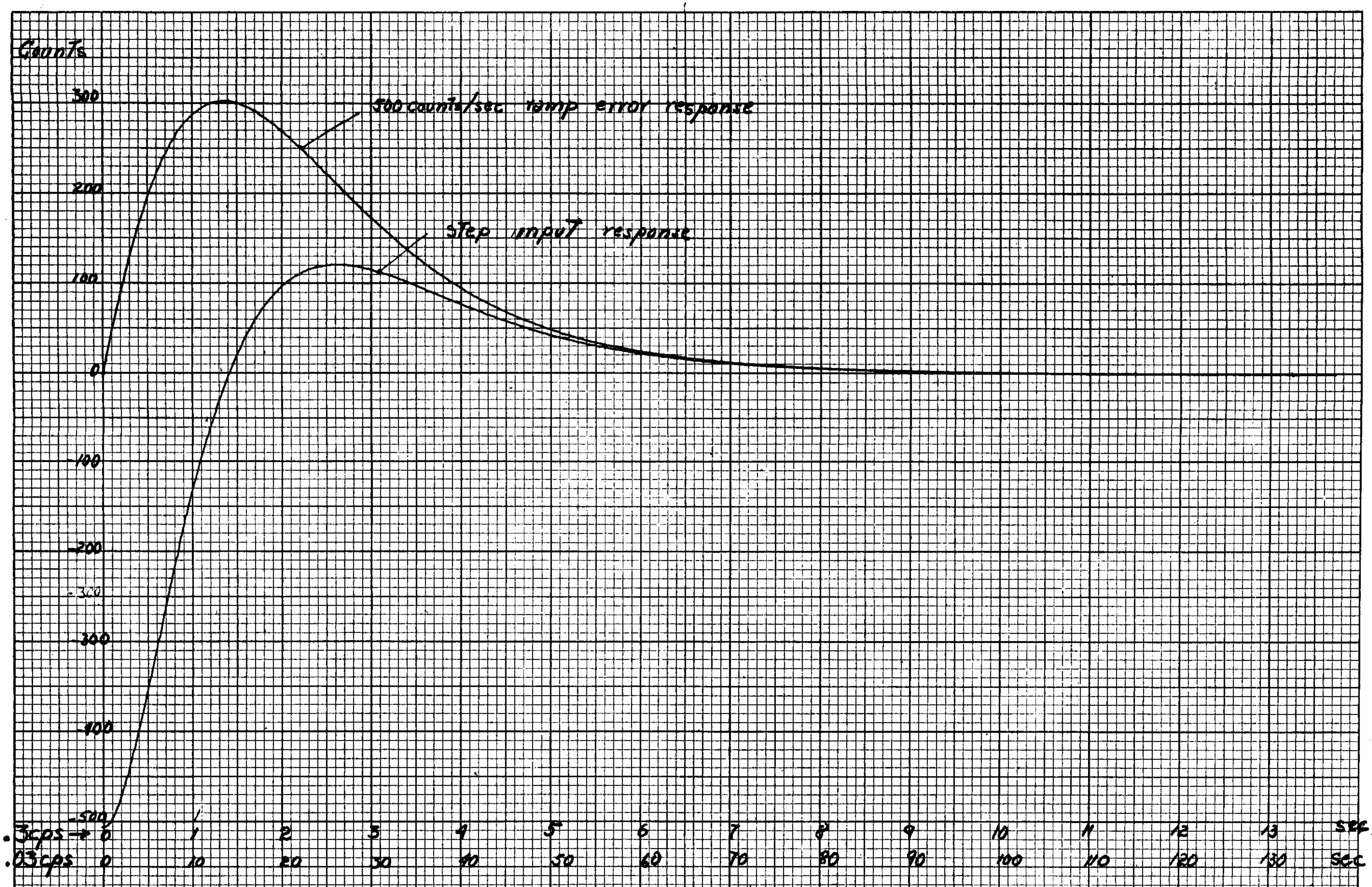


FIG. 2 FREQUENCY RESPONSE



Counts
300
200
100
0
-100
-200
-300
-400
-500
500 counts/sec ramp error response
Step input response
.3cps → 0
1
2
3
4
5
6
7
8
9
10
11
12
13
secs
.03cps
0
10
20
30
40
50
60
70
80
90
100
110
120
130
sec

Figure - 3