ELECTRONIC CONTROL OF A LINEAR SYNCHRONOUS MOTOR

by

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ABSTRACT

Linear synchronous motors with superconducting vehicle windings are being considered as the means of propulsion for high-speed ground vehicles. However, they are highly underdamped with respect to longitudinal oscillations. Damping can be introduced by a feedback control system in which the relative phase of the magnetic fields of the vehicle and the propulsion winding is detected and controlled. The firing angles of thyristors are varied to control the voltage and frequency of current into the propulsion winding. The motor as a whole is a phase-locked loop, with the phase of the armature current locked into the position of maximum thrust.

A 1/25th scale model vehicle has been built with a control circuit based on this principle, and it has been accelerated to 30 mph. This thesis describes the electronic circuits of the scale model motor control system, and discusses the control requirements for synchronization of linear synchronous motors.

THESIS SUPERVISOR: Richard D. Thornton
TITLE: Professor of Electrical Engineering
Acknowledgments

I would like to thank Dr. Richard Thornton for his patience, assistance, and understanding. His ideas created the starting point for this thesis. When I was stuck, his suggestions often broke the impasse. When progress was slow, he was always understanding.

Special thanks are due to Sumner Brown, the senior graduate student on the Magneplane project, who was really the day-to-day supervisor of my thesis. He suggested many of the circuits used in the thesis, and discussed nearly all the circuits and concepts reported in this thesis with me at some length. When I was involved in other projects, he would often work on packaging, construction or debugging of parts of the control system. His loyalty was to the Magneplane project, rather than to his own thesis.

Thanks are due also to Dr. Henry Kolm, the co-creator of the Magneplane concept; Dr. Yuki Iwasa, the most lucid explicator of superconducting phenomena I have ever met; Charles Wallace, for his occasional help with packaging and his continual friendship; Butch Weaver, for help with packaging and circuit construction; Evan Schwartz, for help with packaging; and Christine Plapp, for circuit construction.
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Introduction

It is generally acknowledged that there is a need for an alternative mode of transportation in the medium distance range (25-300) miles in the most densely populated areas of the United States. The highways are congested, dangerous, polluting and comparatively low speed, while the airways in the Northeast Corridor from Boston to Washington are congested to the point of saturation. If a mode of transportation that was high-speed, quiet, nonpolluting and capable of handling heavy traffic densities existed, it would certainly be very welcome. Several new and not-so-new technologies are being tested in the United States and elsewhere, in order to develop a passenger vehicle, capable of providing transportation in the range of 150 to 300 mph. The team of engineers at M.I.T. with which I am affiliated believes that the "Magneplane" concept is the best approach to super-high-speed ground transportation. This concept, an integrated approach to flight which is electromagnetically levitated, propelled and guided, has been described in several review articles (1,2,3), so I will give only a brief description of it.

The Magneplane, as currently conceived, would be propelled by a linear synchronous motor. This motor works by exactly the same mechanism that drives the large, rotating synchronous motors of industry, except that the motor windings are "unfolded" from a cylindrical into a linear geometry. The armature for a full-scale system would consist of a three-phase meander winding of aluminum bars. The field would be supplied by superconducting magnets in the vehicle. Current in the guideway would create a travelling magnetic wave, which would exert a propulsive force on the magnets in the vehicle. Levitation would be provided by aluminum
sheets or bars on the sides of the guideway trough, based on the phenomenon of repulsion between a moving magnet and a stationary conductor. These levitation conductors would also support most of the weight of the vehicle. Guidance is achieved by repulsive forces between the vehicle magnets and the guideway trough. The trough formed by the levitation strips in the Magneplane will be shaped like a V or like the bottom 120° of a cylinder, to permit controlled banking on curves, thus allowing a much smaller radius of curvature for the guideway. Guidance will also be provided by the so-called "keel effect", which occurs if the vehicle magnets' field is strongest in the center of the guideway, and the bulk of the conducting material of the guideway is on the edges, so that the vehicle is strongly centered in the position of least eddy-current formation. This effect should suppress roll, sway and yaw motions of the vehicle. Heave and pitch will probably be suppressed by a trim on the synchronization control of the vehicle, which takes advantage of the fact that the LSM can produce vertical forces of the same order of magnitude as its propulsive force, and that these forces are zero at the maximum thrust position. The vehicle could, of course, also have aerodynamic control surfaces.

The primary advantage of the Magneplane over all other proposed systems of super-high-speed ground transportation is that no part of the system requires a small gap between the vehicle and the guideway. The Magneplane can fly with a naturally stable and soft magnetic suspension at about a foot above the guideway, and does not require the precisely straight, difficult to build and expensive to maintain, track, required by all the "small gap" suspension and propulsion systems.
A Model Magneplane

The Magneplane group at M.I.T. has constructed several 1/25th scale models of a full-scale vehicle, along with several hundred feet of matching guideway. The current vehicle uses samarium-cobalt permanent magnets as the motor field. An iron back-up plate helps hold the magnets in place and provides a return path for the flux. The vehicle has 10 rows of magnets of alternating polarity, with a pole pitch of 3" and a total vehicle weight of 35 pounds. The next vehicle contains three superconducting magnet coils, made from copper-stabilized, niobium-titanium multifilaments wound in 4-layer pancake coils. The pole pitch is 3" and the total vehicle weight is around 40 lb. All tests to date have been run on the samarium-cobalt vehicle. Using the position feedback control system, described in this thesis, this vehicle has been successfully accelerated as a linear synchronous motor at .5 g for 30 feet.

Dr. Yuki Iwasa's field measurements on the cryogenic coils of the superconducting vehicle, indicate that for a given amount of track current, it will accelerate three times as fast as the Sm-Co vehicle. However, because the field of the superconducting vehicle is concentrated much more heavily in the center, where there is no levitation strip, it is not expected that its levitation height at cruising velocity will be significantly higher than that of the Sm-Co vehicle.

The propulsion winding for both the indoor and the outdoor experiments is a flat, two-layer sheet of conducting wires. The two layers are created by a jig which folds an epoxied-flat meander of wires every pole pitch. Current travels the length of the track and returns twice.
The vehicle wheels rest on curved cardboard strips on the indoor guideway. These strips are 45° sections on the side of the track, leaving a middle section of 45° to expose the active length of the propulsion winding. On the outdoor guideway the wheels will rest on PVC tubing, also in 45° sections, until the vehicle reaches the levitation section, where the strips will be curved 3/8" thick sheets of aluminum. The indoor track is 30 feet long. The first outdoor guideway is 400 feet long. It will later be expanded to include a turn. The design and construction of the vehicle and track will be discussed in much greater detail in the doctoral thesis of Sumner Brown.
The Electronic Control System

A block diagram of all the electronics used in the control of the model vehicle is shown in figure 1. A source of 3-phase, 400 cycle power is fed into a cycloconverter. For the indoor experiments, the source of the 400 cycle power was an airplane motor-generator set. The armature of the D.C. motor was supplied with current by a 48 V bank of lead-acid batteries. For the outdoor tests at Raytheon's Wayland division, the power supply is a stiff, 400 cycle source, capable of up to 100 Kw. This voltage is then converted to the desired voltage waveform across the propulsion windings by a cycloconverter, built by Raytheon. For most of our experiments, the cycloconverter controls were set for maximum current, in order to provide the maximum vehicle acceleration. The frequency and phase of the current into the track were controlled by the synchronization system, which is described in this thesis.

The vehicle starts off at standstill, with D.C. current in the track. As the vehicle begins to accelerate, the frequency of the current in the track increases synchronously with the vehicle. To date, we have only accelerated the vehicle to 20 mph, or 30 ft/sec; so with a pole pitch of 3", the current in the guideway has increased synchronously from 0 to 60 Hz. We plan to obtain a speed of around 65 ft/sec or 40 mph, corresponding to a track frequency of 130 Hz or 1/3 of 400 Hz.

The linear motor armature is a Y-connected, 3-phase distributed winding with 20 Cu wires per phase for the indoor guideway on which early experiments have been made and 12 No. 10 aluminum wires per phase for the outdoor windings. The motor winding will be more completely described in Sumner Brown's doctoral thesis. The vehicle field is provided by
ELECTRONIC CONTROL SYSTEM FOR THE MAGNEPLANE

MAGNETIC SENSOR → AMPLIFIER → VOLTAGE CONTROLLED OSCILLATOR

XMTR

RCVR

PHASE-LOCKED LOOP DEMODULATOR

BUTTERWORTH 3-POLE LOW-PASS

DIRECT + INTEGRAL FEEDBACK

VCO

CYCLOCONVERTER SERVO

TO LINEAR MOTOR PHASES

LEAD-ACID BATTERIES → D.C. MOTOR → 400 Hz GENERATOR

FIGURE 1
permanent magnets on the Sm-Co vehicle and by persistent current loops in the superconducting vehicle. A sensor on the vehicle senses the magnetic field of the propulsion windings. The sensor is so placed that, if the current into the windings were a perfectly balanced square wave (or sine wave) the sensor would read zero field when the magnetic field vectors of the armature and the field were aligned in the position of maximum propulsive thrust. The voltage induced in the magnetic sensor, if the propulsion winding currents were sine waves, would be proportional to the displacement from the optimal thrust position for small perturbations. This voltage modulates a subcarrier, which modulates a 42.98 Mc carrier which is transmitted to a stationary receiver, near the power conditioning equipment. The signal is then demodulated, filtered and processed to provide a weighted sum of the phase displacement and the integral of the displacement as the feedback variable of the system. These drive a precision voltage-controlled oscillator (VCO), which creates a square wave, whose frequency is proportional to the feedback signal. This square wave is the input reference signal, which the cycloconverter attempts to duplicate at a higher power level.
The Communications Link

Communication between the vehicle and the ground is achieved with FM-FM modulation and a 1/4-watt transmitter-receiver set. The constraints on the system are that the vehicle electronics should be lightweight, but that the radio link should be powerful enough to overcome the radio frequency harmonics of the magnetic field of the active guideway, as well as the random bursts of radio frequency interference which are indigenous to both the National Magnet Laboratory and Raytheon's Equipment Division at Wayland.

The radio finally chosen was a $60 microphone-transmitter set, which may have been made by Ikefuji of Japan. The transmitter circuit has been deciphered and is shown in appendix II. Apart from its low price, this set has several advantages. The total electronics on the vehicle weighs only slightly more than a pound, which is important because the vehicle must be levitated with a modest amount of force. The transmitting frequency is 42.98 Mhz. This high a frequency is desirable, because the guideway noise associated with SCR transients remains white up to around a megahertz, and then begins to decline as the inverse of the frequency squared. A third desirable feature of this system is that it has an audio bandwidth of 15 Khz, because of its intended use for professional music reproduction. This relatively large bandwidth can be used to frequency multiplex information about longitudinal and vertical position, and, perhaps, other vehicle motions. The transmitter has a useful range of around 120 feet, which has been more than adequate to date, but which may be inadequate when we have longer sections of guideway. At that time, if more power is needed, a VHF, 2 watt marine transceiver can be purchased.
THE COMMUNICATIONS SYSTEM

ON BOARD

HALL PROBE \( V_{out} \)

\( X \) 300 AMPLIFIER \( \text{VCO} \ f_0 = 8.4 \text{ KHz} \)

CURRENT SOURCE

CRYSTAL VCO \( f_0 = 10.745 \text{ MHz} \)

QUADRUPLET AMPLIFIERS

ANTENNA

ON GROUND

RECEIVER

\( 2 \) POLE LOW-PASS \( f_0 = 10 \text{ KHz} \)

LIMITER

PLL DEMODULATOR \( f_0 = 8.4 \text{ KHz} \)

DIFFERENTIAL AMPLIFIER

3-POLE LOW PASS BUTTERWORTH \( f_0 = 80 \text{ KHz} \)

FIGURE 2
A block diagram of the entire communications system is shown in figure 2. Position information from the magnetic sensor is used as a low-frequency voltage to FM-modulate the VCO section of a Signetics 565 phase-locked loop, with a free-running frequency of 8.4 Khz. The output of the VCO FM-modulates the transmitter crystal oscillator. This double wide-band FM modulation provides excellent noise immunity at moderate signal-to-noise ratios. At the receiving end, the microphone receiver retrieves the output of the VCO. This signal is low-pass filtered to eliminate any audio range signals above 10 Khz that might have been added by radio frequency interference at the receiver antenna. The filtered signal is sent through a limiter, whose purpose is fourfold. Weak signals are amplified, strong signals are limited in order not to destroy the input circuits of the 565 phase-locked-loop (PLL), limiting further improves the already good AM rejection of the PLL, and the signal is converted into the form of a square wave, which is the easiest form for the PLL phase detector to interpret. The phase-locked loop generates a signal which has the same frequency and phase as the 8.4 Khz audio carrier. The phase of the carrier and the modulated signal are compared and the instantaneous difference is low-pass filtered to eliminate the carrier and to obtain the original low-frequency signal from the magnetic sensor. The differential output is converted to a single voltage by a differential amplifier. This signal records the magnetic fields from the SCR switching transients fairly faithfully. Since it's undesirable for the frequency of the current to swing wildly every fraction of a cycle and because the only information required is the average relative phase between the vehicle magnets and the magnetic
field of the track, a 3-pole Butterworth low-pass filter was designed to
give an attenuation of 22 db at 400 cycles. Vehicle longitudinal oscil-
lations have a natural frequency of only a few cycles per second. The
3-pole Butterworth has a .707 frequency of 80 Hz. It was therefore
believed, and subsequently shown to be true by experiment, that the
filter would not add much harmful phase shift to the control system at
the response frequencies of interest, while greatly attenuating the 400/sec
switching transients observed at the output of the differential amplifier.
A Brushless D.C. Linear Motor

The first control algorithm that was used to try to accelerate the Sm-Co vehicle was analogous to the method of operation of a conventional D.C. motor. Whenever the position of the motor changed a fraction of a cycle, the position of the field of the current in the track would change a fraction of a cycle. Although this method was discarded, it was not a total failure and a variation on it may be practical for a full-sized vehicle.

The magnetic sensor was placed so that when the vehicle magnets were leading the magnetic field of the propulsion winding by 30 degrees, the sensor would read zero. The relevant waveforms of this system are illustrated in figure 3. In this figure, it is assumed that, because the reference inputs to the servo amplifiers of the cycloconverter are square waves, the current into each track phase is also a square wave. The propulsion winding approximates a uniformly distributed winding, thus the vertical magnetic field of the armature, at an instant in time, is a triangle wave in the longitudinal direction. Because of the phasing of the track currents, there is a travelling magnetic wave in the direction of vehicle motion. The maximum thrust on the vehicle is developed when a north pole of the equivalent vertically pointing magnet on the vehicle is leading the equivalent north pole of the track by 90°. Therefore, the basic control algorithm is as follows. Everytime the vehicle travels 60° or 1°, the polarity of the current in one of the motor phases must change, thereby advancing the field of the armature by 60°. As an example, if the vehicle magnet is 60° in front of the equivalent armature magnet, it will travel until it leads by 120°. At that instant,
WAVEFORMS OF THE LINEAR COMMUTATED
BRUSHLESS MOTOR

MAGNETIC SENSOR OUT AND COMPARATOR IN

COMPARATOR OUT

ONE-SHOT OUT

QA

TO PHASE A REFERENCE

TO PHASE B REFERENCE

TO PHASE C REFERENCE

FIGURE 3
the polarity of the current in one of the motor phases reverses, and the
magnetic field of the armature jumps ahead by 60°, reducing the vehicle
magnet lead to 60°. If this control system worked ideally, the thrust
on the vehicle would oscillate about the maximum thrust at a lead of 90°
and would never fall below $\sqrt{3}/2$ or .866 of the maximum thrust. The best
feature of this system is its simplicity. No linear control analysis is
necessary. The only design problem is to minimize the time delay between
when the magnetic sensor reads zero and when the current reverses in a
motor phase, without destroying the accuracy of each zero-crossing detection.
Unfortunately, in the system we are using, this is impossible, because
of the large discrepancy between the actual and ideal cycloconverter out-
put.

The on-ground circuitry for this system is shown in figure 4. The
vehicle circuits have been explained in the chapter on the communications
link. Immediately before and after the differential amplifier are two
pairs of two-pole low-pass filters, with cut-off frequencies of 1000 Hz.
Their main purpose is to eliminate the 8 KHz carrier of the phase-locked-
loop. It was originally believed that this simple filter would also
eliminate SCR switching transients; but, despite having a repetition rate
of 2400/second, they appeared largely unattenuated at the input to the
comparator.

The output of the filter was ideally identical to the magnetic
field seen by the sensor on-board the vehicle. The normally-low output
of the comparator, then, went high when the field dipped below zero.
However, the comparator tended to change levels more than once during a
zero-crossing. Because of SCR transients, the field seen by the sensor
would go below zero as often as 2,400 times a second when the average voltage was close to zero. To prevent more than one count per zero-crossing from occurring, the output of the comparator triggered a one-shot multivibrator, with a pulse length of 1/1,000th of a second. The length of the pulse was limited by the maximum frequency expected out of the comparator, which was six times one hundred and thirty herz, or 780 cycles per second. The multi-vibrator has a duty cycle of around 98%, making it unlikely that a desired zero-crossing would fail to trigger the multivibrator. The output of the one-shot was then inverted, so that the flip-flop counters would have a falling edge at their trigger inputs immediately as the field went below zero.

The logic sequence of the Q outputs of the three flip-flops is as follows, if all three flip-flops start with an output of Q equals logical zero. On the first zero-crossing, the output of A goes high and then stays high until the fourth pulse. The output of A then stays low until the seventh pulse. The output of B is high from the second to the fifth pulse, while the output of C is high from the third to the sixth pulse. (See figure 5) Thus, the Q output of A corresponds to the A or reference phase of a standard three-phase motor. The Q output of C corresponds to the B or 120° lagging phase, and the not-Q output of B corresponds to the C or 240° lagging motor phase. These logic level pulses are then converted into square waves by the pulse-shaping circuitry. Potentiometer A controls the amplitude A of the square waves. Two operational amplifiers in inverting and non-inverting times-one configurations produce plus and minus A. When the output of A is low, QA is turned off. There is no base drive to QB, and, therefore, no current delivered to the two 10k resistors,
ON-GROUND CIRCUITS FOR THE D.C. BRUSHLESS LINEAR MOTOR - I

PLL DEMODULATOR

LOW PASS

DIFFERENTIAL AMPLIFIER

COMPARATOR

INVERTER

FIGURE 4
ON-GROUND CIRCUITS FOR THE D.C. BRUSHLESS LINEAR MOTOR

FIGURE 5
and no current through the collector of QC. Thus, the operational amplifier in a non-inverting times one configuration is driven by plus-A through a 10K resistor, and gives an output of plus-A. When A is high, the three transistors are turned on, and QC is driven deep into saturation. The three QC transistors have especially low saturation collector-emitter voltages (less than .1V), so that the non-inverting input is connected to minus-A through a much lower impedance than 10K, and the output goes to minus-A. The same process applies to all three outputs. These three outputs become the reference voltages to the three cycloconverter servo sections, which attempt to duplicate those waveforms as higher-level voltages across the motor phases.

Given only occasional SCR noise and outside radio interference, this system should be stable. If the system failed to fire while the vehicle was moving, the vehicle would coast (accelerate for 60°, brake for 180°, then accelerate for 120°) for 1 cycle or 6 inches, and would then create a zero-crossing pulse and lock into synchronism again. If the system received a false pulse at the worst possible time, which is immediately after a desired pulse, the vehicle would travel one-third of a cycle instead of the usual one-sixth before triggering a pulse of the proper phase. In such a case, the instantaneous thrust on the vehicle would remain positive. Low speeds and starting are more of a problem. If the vehicle starts with an angle between the field of the vehicle and the field of the track within the range -60° to 60°, the vehicle will oscillate about the equilibrium zero thrust point at 0°, assuming negligible rolling and bearing friction in the wheels. If the vehicle happens to be placed in the 240° self-starting range, its first oscilla-
tion should swing it past the 60° vehicle lead point in the forward-moving direction and should then trigger the first of a series of synchronizing pulses. Self-starting should be inevitable, because occasional false pulses index the positioning of the magnetic field until the vehicle lies in the 240° self-starting region. It would then require only three correct pulses in a row for the vehicle to store enough kinetic energy to ride out a missed pulse. If the vehicle failed to receive thee correct pulses, the field would continue to index past the stalled vehicle, and it would then try again. Unfortunately, vehicle behavior was not as good as expected.

The experimental vehicle on a thirty foot track accelerated up to around 7 mph. The acceleration was not reliable. The vehicle usually did not self-start, and had to be given a starting push by hand. The main fault with the system was its inability to separate SCR switching transients from the desired field position information.

The electronics was detached from the vehicle, and variable frequency current was fed through the track. With 10 Hz current into the track, the comparator often recorded 4 or 5 zero-crossings for every 10 cycle zero-crossing. It was decided that any attempt to continue making 6 correct switching decisions every cycle unnecessarily magnified the bandwidth of the control system, and consequently increased its susceptibility to noise. In other words, we were creating a system which required a bandwidth of around 1 KHz at cruising speed in order to suppress hunting oscillations, whose natural frequency is predicted to be approximately 2 Hz. The algorithm also assumed a square wave of current into each phase of the track, when the actual current ceases to
resemble a square wave at frequencies as low as 10% of the input alternator frequency. It was decided to switch to the control system described in the next chapter, whereby the phase displacement was controlled directly at low speeds, but held only to an average of zero over several cycles at high speeds.

The conclusions drawn for a magnetic position sensing device, however, do not apply to other position sensing devices. One alternative is a high-frequency magnetic source, which could be used to detect the presence of aluminum bars beneath the vehicle with high reliability. In a full-scale system, even a reliable position sensor would not be exactly analogous to a D.C. motor brush. Some information would have to be stored and processed to compensate for the delay caused by track inductance.
The Control System

The control geometry chosen has the form of a phase-locked loop. An error voltage changes the frequency of a voltage-controlled oscillator, whose phase is compared with that of the input signal. The input frequencies are eliminated by a low-pass filter, which produces a voltage proportional to the difference between the phase of the VCO and the input signal. In a synchronous motor with position, or phase, feedback, the input signal is the true velocity, or mechanical frequency, of the vehicle in radians per second, while the VCO output is the frequency of the current in the propulsion winding. The integral of the difference between these two frequencies is the phase displacement, and this is the output of the magnetic sensor. The 3-pole Butterworth is the low-pass filter of the phase-locked loop, and the filter's output is the error voltage which controls the frequency of the current into the track.

The phase-locked loop method of control is substantially different from the usual method of motor speed control by control of the armature voltage. It is more precise, because there is a one-to-one correspondence between armature current frequency and motor speed in a synchronous motor, while the ratio of terminal voltage to back e.m.f. or speed changes with the velocity in any motor. Also, because the average energy storage per cycle doesn't have to change when only the frequency is being controlled, the time constant associated with the control system is smaller than that for voltage control. Unfortunately, the full advantages of phase-locked loop control are not attained by this motor, because of the difficulty of obtaining a precise measurement of
the phase of the current into the propulsion windings.

The cycloconverter can be thought of as a set of 18 switches, switching selected parts of a 400 Hz, 3-phase input signal into 3 motor phases in order to form a gross approximation to a desired low frequency signal in the three motor phases. In a large motor, the SCR switching transients would be filtered out by the large inductance of the motor; but with the geometry we are using, the propulsion winding is primarily resistive, with a Q of only 1 at 500 Hz. Therefore, SCR switching transients are faithfully picked up by the magnetic sensor, with a repetition rate corresponding to 400 Hz. A 3-pole Butterworth filter with a cutoff frequency of 80 Hz was installed between the sensor and feedback VCO, in order to give an attenuation of at least 22 db at 400 Hz. Since the natural frequency of longitudinal oscillation is in the neighborhood of 2 cycles per second, it was expected that the phase propagation delay introduced by the filter would not prevent accurate control.

The specification that had to be met by the control system was that the vehicle had to maintain synchronization starting from standstill, through acceleration up to \(5\text{m/sec}^2\) during perturbations in guideway alignment and winding distribution. No specifications were made concerning "ride quality", because the first objective was to accelerate the vehicle as rapidly as possible to a velocity at which it could maintain its velocity when levitated.

In order to achieve rapid and approximately critically damped response, conventional trial-and-error root-locus techniques were used. (4) Information on the relative position, or phase, and the integral of the
position of the vehicle and armature fields are fed back. The direct feedback changes the phase of the current rapidly in order to follow sudden perturbations in total thrust or drag and is also used for starting, to avoid integrator saturation. The integral feedback is necessary to keep the average long-term position error equal to zero. Neglecting dynamic considerations, the gain of the integral feedback loop should be as high as possible, because it is responsible for the major objectives of the system: to keep the position error finite during acceleration and to keep the average position error equal to zero during steady-state cruising. However, increasing the integral feedback increases the open loop phase shift and decreases the damping of the system. Therefore, the basic design problem is simply to find the ratio of integral to direct feedback gain that allows the maximum value of integral feedback gain without making the system underdamped. The trial-and-error results are presented in figures 6-10.

The root-locus of this system is typical of many servomechanisms. The open-loop zero at \( s=0 \) prevents the dominant pole from crossing the imaginary axis. The other poles leave the real axis near the filter cutoff frequency, nearly 2 orders of magnitude farther away from the imaginary axis than the dominant pole created by the ratio between direct and integral feedback. This configuration will stand very large amounts of feedback, even with the Butterworth modeled with all 3 poles (see figure 11), before it goes unstable.

The circuit used to realize this root locus involves a few modifications of the previous D.C. linear motor. The comparator has been replaced by a 3-phase Butterworth filter, a weighted sum of an integrator
Block Diagram of the Control System for the Experimental Sm-Co Vehicle

\[ \frac{K_{FE}}{mS} \rightarrow \frac{1}{S} \rightarrow V_{t} \rightarrow \frac{1}{S} \rightarrow X_{d} \]

\[ \frac{\beta}{S+\beta} \]

\[ K \]

\[ k \]

\( K_{FI} \) = the motor gain constant

\( m \) = mass of the vehicle

\( \beta = 0.707 \) frequency of the low pass filter

\( K \) = loop gain of direct feedback path

\( k \) = loop gain of integral feedback path

\( K_{FV} \) = velocity/voltage; gain of the VCO multiplied by the motor pole pitch

Evaluate the fixed constants:

Gain of 8038 VCO is 770 Hz/3.4V = 228 Hz/V
Frequency into track = 228/6 = 38 Hz/V
Velocity gain = 6" x 38 = 0.15m x 38 Hz/V = 5.7 \( \frac{m}{sec} \)/Volt therefore

\( K_{FV} = 5.7 \)

Output of Butterworth at max. field = 0.7V therefore
Gain of voltage/m displacement = \( \frac{2\pi}{\lambda} \) \( \cdot \) 0.7V = \( \frac{2\pi}{0.15} \) \( \cdot \) 0.7V = 29V/m

Since this doesn't appear explicitly in the model, divide desired \( k \) and \( K \) by 29 to get gains of the operational amplifier integrator and amplifier, respectively.

FIGURE 6
Roots of the Experimental System

a. without including the low-pass filter in the model

\[
\frac{X_d}{V_t} = \frac{1/s}{1 + \left(\frac{1}{s}\right)K_{fv}K + \frac{K_{fv}k}{s}} = \frac{s}{s^2 + K_{fv}Ks + K_{fv}k}
\]

For experimental system, \(K = (0.5)(29) = 14.5\)
\(k = (5)(29) = 145\)
\(K_{fv} = 5.7\)

therefore
\[
\frac{X_d}{V_t} = \frac{s}{s^2 + (5.7)(14.5)s + (5.7)(145)} = \frac{s}{s^2 + 835 + 827}
\]

\[
= \frac{s}{(s+71)(s+11.6)}
\]

The system is overdamped.
Roots of the Experimental System with a single-pole filter in the model

\[ \frac{X_d}{V_t} = \frac{G}{1+GH} = \frac{s(s+\beta)}{s^2 + s^2 + \beta K_{fv} K_s + K_{fv} K} \]

\[ \frac{X_d}{V_t} = \frac{G}{1+GH} = \frac{s(s+\beta)}{s^2 + 360s^2 + 30,990s + 309,900} = \frac{s(s+360)}{(s+10)(s+350s+30,990)} \]

For the experimental system, \( \beta = 360 \), \( K_{fv} = 5.7 \), \( K = 14.5 \), \( k = 145 \)

For the second-order poles \( \omega_n = \text{natural frequency} = 176 \), \( \zeta = \text{damping coefficient} = .98 \)

Therefore the model system is critically damped. Other predicted performance includes:

For an acceleration of .5g, \( X_{dss} = \frac{a}{K_{fv} k} = \frac{4.9m/sec^2}{(5.7m/sec/V)(145V/m.sec)} = .6cm = 14^\circ \)

For an acceleration of 1.5g on cryogenic vehicle

\( X_{dss} = \frac{a}{K_{fc} k} = \frac{1.5}{(5.7)(145)} = 1.8cm = 42^\circ \)

For a velocity step change of 1m/sec

\( \Delta X_d = \frac{\Delta V}{K_{fv} k} = \frac{1}{(5.7)(145)} = 1.2cm = 28^\circ \)
ROOT LOCUS OF THE EXPERIMENTAL SYSTEM
WITH THE OPEN-LOOP GAIN VARIABLES AND THE
RATIO OF INTEGRAL TO DIRECT FEEDBACK GAIN = 10

A. NO FILTER

B. WITH ONE-POLE FILTER, $\omega_o = 360$

FIGURE 9
Stability of the Experimental System Using an Exact Model of the Butterworth Filter

\[
\frac{X_d}{V_t} = \frac{G}{1+GH} = \frac{1}{s[1 + \frac{5.7}{s}(14.5 + \frac{145}{s})] (1.25\times10^8/s^3 + 10^3s^2 + 5\times10^5s + 1.25\times10^8)}
\]

\[
= \frac{8\times10^{-9}s^5 + 10^{-3}s^4 + 5\times10^5s^3 + 1.25\times10^8s^2 + (5.7)(14.5)s + (5.7)(145)}{s^5 + 10^3s^4 + 5\times10^5s^3 + 1.25\times10^8s^2 + (5.7)(14.5)s + (5.7)(145)}
\]

Determine stability by Routh's Criterion

| $s^5$ | 1 | 5\times10^5 | 83 |
| $s^4$ | 10^3 | 1.25\times10^8 | 827 |
| $s^3$ | 3.75\times10^5 | 82 |
| $s^2$ | 1.25\times10^8 | 827 |
| $s^1$ | 82 | 0 |
| $s^0$ | 827 |

All entries in the first column are positive, so the system is stable.

FIGURE 10
and an amplifier, and a precision VCO. (See Figure 11). The input of the Butterworth filter is a fairly accurate reproduction of the magnetic field seen by the vehicle sensor, along with a few hundred millivolts of the Signetics 565 carrier signal. The three-pole Butterworth effectively removes all components of the signal with frequencies above 80 Hz. In particular, 400 Hz switching is attenuated by 22 db. At cruising speed, this means that the propagation delay of the control system will be equal to a couple of cycles. A rough estimate of the natural frequency of longitudinal oscillation can be obtained from the measurement that the 90° thrust angle acceleration of the vehicle equals 5m/sec², while the thrust at 0°, 1.5 inches (3.8cm) away, equals zero. Therefore, the stiffness of the linearized equivalent spring is 5m/sec²/.038m = 128 sec⁻², corresponding to a vibration frequency of $\sqrt{128}/2\pi = 1.8$ Hz. At the 90° thrust angle, the equivalent spring stiffness is zero, so the natural longitudinal vibrational frequency should be even lower than 2 Hz. Thus, the control system should have an excess of bandwidth for reacting to change in the system.

The output of the Darlington filter is then fed into a times minus 1 inverting amplifier and into an integrator with a 1/RC gain of $-10$ sec⁻¹. The outputs of the amplifier and the integrator are summed by a times 1/2 inverting amplifier. Thus, the integrating section has a gain of 5, while the non-integrating section has a gain of .5. These can be shown to correspond to the desired open-loop gains called for by the root-locus analysis by two measurements on the system. With full current into the propulsion winding, the vehicle electronics was detached from the vehicle and the magnetic sensor was moved longitudinally until a maximum
THE CONTROL CIRCUIT

LOW-PASS

BOARD A

PLL FM DEMODULATOR

DIFFERENTIAL AMPLIFIER

DARLINGTON 3-POLE LOW-PASS

INTEGRAL FEEDBACK

DIRECT FEEDBACK

FIGURE 11
was observed at the output of the Butterworth filter. The maximum voltage observed was .7V, corresponding to an incremental gain, for a winding wavelength of 6" (15.2cm), from the phase displacement to the input to the control operational amplifiers of $2\pi \times 0.7V/15cm = 29V/m$. (see figure 8). The input to the 8038 precision VCO was lowered by 3.4V, and its frequency increased by 770Hz, for a gain of 228 Hz/volt, or a synchronous vehicle velocity gain of 15 cm times the frequency, or 5.7 (m/sec)/volt. The desired direct open loop gain is 827. Thus, the gain of two series operational amplifiers should be $82.7/(29)(5.7) = .5$, while the gain of the series integrator and adder should be $827/(29)(5.7) = 5$.

The integrator also has a switch and a 2.2 megohm bleeder resistor across the integrating capacitors. The bleeder resistor will prevent offset error from accumulating for longer than around 7 seconds. The switch is to prevent capacitor saturation during start up. At the start, the switch is closed, and the operational amplifier output is near zero. The vehicle is accelerated for a few feet using only direct position feedback, then the switch is opened. Since the vehicle is synchronized, the input into the integrator will stay low, and the integrator will not saturate. The timing for throwing the switch is not critical. The switch has been thrown after only a few inches of travel, and the vehicle invariably locks into sync. If one waits too long, the vehicle ceases to accelerate, but usually does not lose synchronization. The trimpot at the non-inverting input to the adder makes self-starting more probable, because a slightly less than zero input voltage is needed to zero the output frequency of the 8038.
The trimpot is adjusted so that the output of the 8038 is just barely still D.C. A diode connects the input of the VCO to ground in order to protect the I.C. from large positive voltages. The 8038 also has the property that, if the input voltage goes more negative than around 7V, the output will change rapidly from maximum frequency to D.C. A 6.2V Zener clamps the output at a maximum of 1KHz.

The switch at the input to the 8038 VCO disconnects the feedback control system and allows the frequency of the current in the track to be varied by hand. The linear synchronous motor was accelerated by "eye feedback" before the automatic feedback control system was completed. Hunting was visible in every cycle, and no speed greater than 1 or 2m/sec was ever achieved. The capability of running the vehicle by hand is retained for the purpose of demonstrating the superiority of automatic feedback control, and for experimental convenience.

The present control system is also expected to be self-starting, self-velocity-limiting at cruise velocity, and capable of regaining synchronization after a gross disturbance.

Behavior at starting is best understood by dividing the relative position of the vehicle and armature fields into four regions: A, B, C and D, as shown in figure 12. In regions C and D, and VCO is "on", or, stated otherwise, in regions A and B, the output polarity of the VCO, and, therefore, the current into the propulsion windings does not change. If the system is in state A, the motor moves forward until it hits the boundary between A and D, which is the maximum thrust position. The vehicle continues to be thrust forward; but, after an undetermined but short delay, the VCO output polarity changes. The position of the
BEHAVIOR STARTING FROM STANDSTILL

A B C D

FIELD SEEN BY THE SENSOR

NORMAL OPERATING POINT

MAXIMUM THRUST

FORCE ON THE VEHICLE

DIRECTION OF MOTION

FIGURE 12
armature field advances 60° and the system is back in Section A. This is the normal mode of operation. Note that the mechanism of propulsion control is identical to that of the earlier D.C. linear motor. If the vehicle starts in position D, it goes forward. The VCO polarity changes. The armature field advances, and the polarity keeps changing until the system is back in A. If the vehicle can advance into C before the current in the track switches, or if it starts in C, the vehicle begins to be accelerated backward. If there were no VCO, it would oscillate around the stable equilibrium point at the boundary between C and D; but the VCO changes polarity in both C and D, and the armature field keeps advancing until the system is in A. If the system starts in B, two paths are possible. Sometimes the vehicle will move backward into C, continue moving backward into D, slow down in D, and then the continually indexing VCO will place it in A, where it will reverse velocity, and start off in the normal forward mode. The other possibility, especially likely if the system starts out near the A boundary, is that the vehicle will pick up a fair amount of momentum in B; it will then accelerate more in C than it is decelerated in D, so that the sum of $P_B + P_C - P_D$ will be greater than the inevitable positive increment in momentum as it traverses A. The system is then in B again, has some accumulated negative momentum and continues backward. If rolling and bearing friction were zero, the average cycle thereafter would produce zero net momentum, but there might be several "negative" cycles in a row. Since there is drag in the system, one would expect the system to eventually slow down and reverse directions. Within our experience, the experimental vehicle starts off in the wrong direction approximately
one-eighth of the time. It could easily be prevented from doing so by an optical sensor and two stripes at the starting position, or by other means. But, at present, we have not solved the problem of 100% sure self-starting, using only magnetic field sensing.

The vehicle is expected to limit its velocity and go into a steady-speed cruising mode at the velocity determined by the Zener diode across the 8038 input, approximately 60 mph. (Of course, the vehicle may be velocity limited before that, as the output waveform of the cycloconverter becomes less and less ideal at higher speeds, the average force may diminish until it is equal to the sum of the magnetic and aerodynamic drag.) As the vehicle goes faster than the limiting velocity, the VCO output clamps at the cruising frequency, the vehicle lead increases, the thrust angle advances to zero, the propulsive force on the vehicle decreases, and the vehicle returns to below the clamping velocity. The cycle then repeats itself.

Finally, within a range which has not yet been determined, if the system loses synchronization, it can regain it. As the field of the track currents slips in respect to the vehicle or vice-versa, the output range of the 8038 is swept through the entire range of possible frequencies into the track. If the slip is not excessive and the VCO is not sweeping too fast, when it goes through the frequency corresponding to the vehicle velocity, the vehicle will lock into synchronism. I haven't analyzed the problem, but it seems possible that since the vehicle can accelerate at $5\text{m/sec}^2$, corresponding to 32 cycles/sec/sec, that, if it has 180°, with an average acceleration of $2/\pi$ times $5\text{m/sec}^2$ to pull into sync., it should survive a disturbance which caused it to
slip at 8 cycles/sec for a second and still be able to pull into sync. For example, when the input to the 8038 was misadjusted, so that the zero input signal produced 30 Hz out, corresponding to 5 Hz into the track, the vehicle still self-started half the time.

If magnetic sensing is used for a full-scale vehicle, the type of control system used for the experimental vehicle should be entirely adequate. The ratio between the characteristic mechanical time constant of the vehicle flight to the electrical time constant of the control should increase by a factor of 10, giving the control system additional time for filtering or data processing. The much higher L/R ratio of a full scale system will produce track currents with far less harmonic distortion than that of the present system. This will produce greater motor efficiency as well as relaxing the requirements for further filtering in the control system. A third possible advantage is that it will probably be more economical to rectify line power and invert it to form controlled frequency three-phase power, than to use an on-site source of higher than 60 cycle power to be cycloconverted. An inverter should be advantageous for control, because the distortion of the waveform will not increase greatly as the vehicle approaches cruising velocity.

It is my feeling that synchronization control for a large-scale system will be so fast and accurate that it can be ignored in preliminary studies of such a system.
The Heave Control Circuits

The basis of Magneplane's heave damping is to apply vertical forces to the vehicle by changing the phase angle between the magnetic fields of the vehicle and the armature.

The circuitry which will be used to suppress heave motions in the model vehicle has been reported in the bachelor's thesis of John Johl. A Setra 0-1 g accelerometer is positioned in the vehicle to measure vertical acceleration. After being amplified, its output is used to modulate the VCO section of a Signetics 565 PLL, exactly as was done for the synchronization by position sensing system. The free-running frequency of the vertical acceleration sensing VCO, however, is 3.9 KHz, or slightly under half that of the longitudinal position sensing VCO, in order to prevent the possibility of the 8.4 KHz receiver PLL locking on the second harmonic of the wrong signal. The two VCO outputs are multiplexed by a simple resistive adder, and then sent through the transmitter.

At the receiver end, the signal is filtered, then demultiplexed and demodulated by phase-locked loops, with free-running frequencies of 8.4 and 3.9 KHz, respectively. The output of the 3.9 KHz PLL is not sent through a Butterworth filter, because, after a modest amount of 500 Hz center frequency filtering to eliminate the PLL carrier, there should be no significant sources of noise in the system reporting vertical acceleration information. The signal will be integrated, in order to create a voltage proportional to the vertical velocity of the vehicle. This voltage will be sharply limited at a voltage corresponding to a motor phase shift of $\pi/8$ degrees. Thus, if the heave control circuit contributes
a worst case phase shift of \( \pi/8 \) and the synchronization control lags by a shift of \( \pi/8 \) degrees in the same direction, the total thrust on the vehicle will only be diminished by a factor of 2. Probably velocity feedback alone will be used, with no direct acceleration feedback. As the block diagram and equations in figure 13-14 show, the main effect of acceleration feedback is to decrease the natural frequency of oscillation, while the major effect of velocity feedback is to add damping to the system. Damping is needed desperately, but the control system ought already to be fast enough to handle the oscillation frequency of 7 cycles/sec predicted for the heave motion. Because the control power available is limited by the requirement that synchronization must be maintained, acceleration feedback is a luxury for this model.

As shown in the equations of figure 17, with the system parameters currently expected, the control system will saturate, if a damping factor of 0.7 is required, at a vertical velocity of only 2.9cm/sec. John Johl, in his thesis, prevents saturation of the system by allowing a Q factor of about 5. On reconsideration of the problem, however, it is clear that there is no advantage to be had in insisting that the control be linear. The system should be allowed to saturate much of the time and use all the control force allowed to it to diminish the stored kinetic potential energy of the vehicle. Therefore, the model will have a velocity gain constant of at least 60, or a damping constant of at least 0.7 in the linear region.

There may be considerable error in the estimate of the equivalent suspension spring constant, but the overall conclusions remain the same. Even if the suspension force is only proportional to the inverse height,
Block Diagram of Heave Control System

\( h_{\text{des}} \) is the equilibrium suspension height. It does not have to be specified beforehand as the absolute distance between the guideway and the vehicle. However, in order to calculate the equivalent spring constant of the suspension, the absolute distance is assumed to be a levitation height of 1/2 inch. For the control equations, initial \( h_{\text{des}} \) is assumed to be zero. For a sudden guideway step of 1 cm, \( h_{\text{des}} = 1 \text{ cm} \delta_{-1}(t) \).

\( h_{\text{true}} \) is the actual height of the vehicle above the track, with the convention that \( h=0 \) is the initial equilibrium position.

\( K \) is the incremental spring constant of the magnetic suspension = -36,000 N/m.

\( m \) is the mass of the vehicle = 15 Kg for the Sm-Co vehicle.

\( k_a \) is the acceleration feedback gain constant.

\( k_v \) is the velocity feedback gain constant.

Figure 13
Heave Damping Control

\[ h_{\text{true}} = \frac{G}{1+G} = \frac{K/(1+k_a)M}{s^2 + \frac{k_v}{(1+k_a)} s + \frac{K/m}{(1+k_a)}} \]

Expected thrust on vehicle = 0.6g = maximum possible heave force. Heave control is limited to an excursion of \( \pi/8 \), therefore the maximum control heave force, corresponds to an acceleration of 0.6sin \( \pi/8 \) = 0.2g = 2m/sec^2.

\[ K = 27,700 \text{ N/m}, \quad M = 15Kg \]

\[ h_{\text{act}} = \frac{(27,700)/(1+k_a)15}{s^2 + \frac{k_v}{(1+k_a)} s + \frac{2,400}{(1+k_a)}} \]

Now \( \omega_n = \frac{K/m}{(1+k_a)^{1/2}} \) if \( k_a = 0 \), \( \omega_n = 43 \), \( f_n = 7 \text{ Hz} \).

\[ \zeta = \frac{k_v}{2(1+k_a)^{1/2}} (K/m)^{1/2} \]

If \( k_a = 0 \), \( \zeta = \frac{k_v}{2\omega_n} = \frac{k_v}{86} \)

If \( k_v \) is chosen to = 60.8, \( \zeta = 0.717 \), so that damping will be reasonably high, then the control system will limit at \( k_v \cdot 0.2g \)

\[ v = 2/60.8 = 3.3 \text{ cm/sec} \]

Figure 14
the equivalent spring constant decreases only by a factor of two. If the suspension height is 1 inch, instead of a 1/2 inch, the spring constant will be down by a factor of eight, but the motor force will be halved, which is unacceptable. The saturation velocity of 3 cm/sec corresponds to only a few thousandths of a second of free-fall. The methods currently used to adjust the guideway height cannot distinguish between differences of a few millimeters. It appears almost inevitable that the heave damping system will be a bang-bang control system.
The Block-Switching Circuit

The 1/25th scale models on which we are conducting our experiments will eventually be tested on a guideway as long as 1400 feet, in order to give the vehicle adequate time to accelerate to cruising speed, to cruise stably at its steady-state velocity and to decelerate. The sources of variable electric power easily available to us are incapable of providing the power necessary to excite 1400 feet of propulsion windings. Therefore, it is necessary to switch power from one block of windings to another, as the vehicle passes over the junction between them.

A rough estimate of how much propulsion winding could be excited by our present system was obtained using Dr. Yukikazu Iwasa's plots of the field strength under the vehicle cryogenic coils. The voltage across any one phase is limited to 100 V maximum by the reverse voltage capabilities of the cycloconverter SCRs. The track winding has a measured resistance of .05 Ohm/10 foot section (3.1m). The force necessary to propel the Sm-Co vehicle at cruising speed is obtained from towing experiments, which indicate that the magnetic drag at 45 mph (20m/sec) is equal to 5m/sec². The available force from the motor is obtained from the three-phase, synchronous motor equation \[ F = 1.5n B_{\text{max}} I_p w \sin \alpha, \] (5) where \( n \) is the number of alternating polarity magnets on the vehicle, \( B_{\text{max}} \) is the maximum flux density of the vehicle magnets seen at the track windings, \( I_p \) is the peak current per phase in the track winding, \( w \) is the effective width of the track winding and \( \alpha \) is the motor torque or thrust angle. The thrust angle is assumed to equal 90°. The active length of a guideway coil is 3 in. (7.6cm). An effective length of 4 inches (10.2cm) is assumed. The maximum field from the Sm-Co magnets at a levitation height
of 1cm equals 2 Kgauss. Therefore, for a thrust of 5m/sec^2 a peak current of 140 amps, or an rms current of 100 amps is necessary for the Sm-Co vehicle. A rough integration of Dr. Iwasa's charts of the flux density beneath the cryogenic vehicle was used to obtain an average flux density. The net result was that a current of only 35A rms should be necessary to provide 5m/sec^2 of thrust to the cryogenic vehicle. The magnetic drag on the cryogenic vehicle at cruising is assumed to be 5m/sec^2 as there is no reason to expect a drastic change from the measured lift-to-drag ratio of the Sm-Co vehicle with the same guideway and a similarly shaped vehicle.

If 35A rms are necessary to maintain steady state, then, neglecting back e.m.f., the longest section of propulsion winding that could be excited at 100V, the rated output voltage of the cycloconverter, equals length X .05 Ohm/10 ft. (3.1m) X 35 A rms = 100 V rms, or length = 570 feet (175m). The actual length that can be excited will be less than 570 feet (175m), because there will be a back e.m.f. equal to K, the motor constant = .73 V-sec/m times the velocity = 22m/sec, or 16V. Also, the length that can be excited will be reduced, because the waveform into the track at high speeds is pulsed and has a high ratio of peak to r.m.s. voltage. Therefore, while it is possible that three blocks of around 500 feet (153m) will be adequate for a 1400 foot (420m) track, we have designed block switching for at least four blocks.

The block switching geometry chosen is shown in figure I-A. At any one time, only one of the switches is closed, and current flows through its associated block. When the vehicle is about to reach the intersection between two blocks, the contactor associated with the forward
block closes. The two sections are thrown in parallel across the cyclo-converter voltage source, so the current remains approximately constant. When the vehicle is entirely onto the forward block, the switch to the rear block is opened. A smoother transfer of current could be achieved by the use of the configuration shown in figure I-B, in which all but one of the switches are closed, the forward switch opens at an intersection,

![Block Switching schematic](image)

A. - Block Switching for the Experimental Propulsion Windings

B. - Suggested Block Switching for a Full-Scale System.

Figure 1

then the rear switch closes. This configuration would probably be used in a full-scale system, to insure passenger comfort. It was rejected for the model, because of the danger of putting a dead short across the cycloconverter, if any switch should ever fail to open.

The logic for block-switching is divided into two parts (see figure 15). The decision to switch from the first block to the second is made with a light source and a photoresistive element, which are stationed at the intersection of the two sections. During its passage through the second and all subsequent blocks, the vehicle must be in synchronization,
THE BLOCK SWITCHING SYSTEM

LIGHT SOURCE

BLOCK 2 → VEHICLE → BLOCK 1

CYCLO CONVERTER CONTROL INPUT

PHOTODETECTOR

ONE-SHOT

FLIP FLOP

BEGIN COUNT

FLIPS COUNT FROM 1 TO 2 WHEN VEHICLE PASSES

COUNTING LOGIC

DECADe COUNTERS

COUNT TO FOUR FLIP-FLOPS

4-OUTPUT LOGIC

2 A RELAYS

3-Φ, 40 A RELAYS → 4 BLOCKS OF WINDINGS

CYCLO CONVERTER OUT

FIGURE 15
so the distance it travels can be determined to within six inches, simply by counting cycles. Therefore, as soon as the vehicle has switched to block two, a programmable counter begins to count cycles of the control input into phase A of the cycloconverter. It then counts up to a pre-programmed number, corresponding to the length of the block, then switches current to the next block.

The logic circuitry is shown in figure 16. There are two output flip-flops, whose four possible states correspond to the four blocks being switched. At the beginning of a run, the \( C_D \) asynchronous input of both flip-flops is low, so the \( Q \) output is 0 0. When the photodetector detects reflected light, its resistance decreases sharply by at least half for any non-pathological vehicle trajectory. The second-stage transistor saturates, removes the asynchronous restraint on the flip-flops, then triggers them. The output of flip-flop A goes high. Before the vehicle was detected, the logical expression for the flip-flop outputs was \( \overline{A} \cdot \overline{B} \). Therefore, of the four NAND gates connected to the flip-flop outputs, only the one connected to \( \overline{A} \) and \( \overline{B} \) will have a low output. This level is inverted by a transistor driving the coil of a 2A relay. The 2A normally-open relay closes, applying 120 VAC wall power to the exciting coil of a 40A, 3-phase, contactor in series with the cycloconverter and the first block of track. When flip-flop A goes high, excitation is removed from the block 1 contactor and is applied to the block 2 contactor through the \( A \cdot \overline{B} \) NAND gate.

At the same time that the first block is switched, input 3 of the decade counters goes low. This removes an asynchronous constraint that the output of all the decade counters should equal zero, and the output
of the series decade counters indexes one, each time the output of the
phase A control input for the cycloconverter goes low. The eight most
significant bits from the two most significant decade counters go into
a counting logic block whose output is not - (1 and 2 and 3 and 4 and
5 and 6 and 7 and 8). In other words, when all eight inputs from the
decade counter are high, the logic output goes low. The logical inverse
of each output is also produced in this block, so that the block length
can be programmed to intervals of 10 feet, and then the desired com-
bination of decade counter outputs and inverse outputs is wired into the
eight inputs of the "not-all-of-these" logic circuit. When the desired
number of 10 foot lengths have been traversed, the logic counter goes
low. This generates a pulse by a R-C high-pass filter. This pulse sets
the output of all three decade counters to zero. It also causes a
trigger to the 2 flip-flops. The Block 3 relay is closed and the
Block 2 relay is opened. A similar sequence of events precedes the
transfer of current from Block 3 to Block 4.

The low-level circuitry has been built by Christine Plapp.
Control of a Full-Scale System

A full-scale system will have more position information and more time to process that information than does the 1/25th scale model. But, in many ways, the control requirements on it are far more stringent. A full-scale system will not only have to remain in synchronism, but it will have to provide a ride which is comfortable enough to gain passenger acceptance. It will probably also have to meet federal ride-quality standards. Furthermore, loss of synchronism is not tolerable in a full-scale system. While it would not cause physical injury, it could cause dozens of miles of the system to be shut down until the vehicle could be removed.

There should be essentially four modes of operation in a one-way trip on a full-scale vehicle: taxiing, acceleration with levitation, cruising and braking. Low-speed acceleration on a non-conducting guideway is less difficult to achieve than high-speed acceleration with levitation. A full-scale vehicle might enter a levitation strip at a speed of around 60 mph, and then accelerate in flight to over 200 mph. During this time, it could be controlled by a cycloconverter, as was our experimental model. More probably, it would be controlled by a rectifier-inverter set, eliminating the need for 6 megawatt, 400 cycle power. The control equations would be similar for either type of system. At cruising, the vehicle will probably be run from 60 cycle line power. Active control will be provided by control of SCR firing angles, but there will be no direct control of frequency. Braking control will be similar to acceleration, with active control of voltage and frequency. Braking will be regenerative, and passive braking is also
inevitable, through aerodynamic and magnetic drag. Controlling deceleration is similar to controlling acceleration, but involves the need for little or no control power and is ignored here.

Frequency control, during acceleration and deceleration, can probably be accomplished so rapidly that it should be ignored in a first-order analysis of the system. The propagation delay of the communications link should be no greater than that of the experimental vehicle. Filtering requirements should be less, because the armature current will be much more sinusoidal than that of the model. The L/R of the propulsion winding can be ignored, because the frequency of a current into an inductor can be changed instantaneously, while the phase can be changed within a fraction of a cycle. At the same time, the period of longitudinal oscillation will increase from around 1/2 second to something ranging from four to six seconds, depending on vehicle velocity. However, there will be a time delay in the system if cycloconverters are employed, because they can only act within a significant fraction of a cycle of their power supply. A worst case analysis was done of a system with a pure time delay of 1/240 of a second, corresponding to a quarter of a cycle delay with a 60-cycle source. Using the approximation $e^{-Ts} = -(T/2)s + 1/(T/2)s + 1$, integral and direct position feedback gain constants were found, such that the synchronization system had a damping coefficient of .8, with a decay time constant of 1/30th of a second (see figure 17). This is so rapid that the synchronization system can be ignored when considering acceleration.

The parameters used in the analysis of a full-scale system come from an internal memorandum issued by the Raytheon Advanced Development
Synchronization System With A Time Delay

\[ V_t \xrightarrow{+} \Sigma \xrightarrow{+} \frac{1}{s} \]

\[ e^{-Ts} \xrightarrow{+} K \xrightarrow{+} \frac{k}{s} \]

\[ K = \text{open-loop direct gain} \]
\[ k = \text{open-loop integral gain} \]
\[ T = \text{time delay} \]
\[ e^{-Ts} \approx \frac{-(T/2)s + 1}{(T/2)s + 1} \]

\[ \frac{X_d}{V_t} = \frac{G}{1+GH} \text{ where } G = \frac{1}{s}, H = (K + \frac{k}{s}) \frac{-(T/2)s + 1}{(T/2)s + 1} \]

\[ = \frac{s(s+2/T)}{s^3 + \frac{2}{T} - ks^2 + \frac{2K}{T} - k} \]

Let \( K = 50, k = 1,000, T = \frac{1}{4} \times \frac{1}{60} \text{ sec} = \frac{1}{240} \text{ sec}. \)

\[ \frac{X_d}{V_t} = \frac{s(s+480)}{s^3 + 430s^2 + 23,000s + 480,000} = \frac{s(s+480)}{(s+370)(s+30-j20)(s+30+j20)} \]

Damping constant \( \zeta = .83; \)

Natural frequency \( f_n = 6 \text{Hz} \)

Position error for an acceleration of \( .1 \text{g} \)

\[ X_{\text{ds.s.}} = sX|_{s\rightarrow0} = \frac{480(.1 \text{m/sec}^2)}{(480 \text{sec}^{-1})(1,000 \text{sec}^{-1})} = .1 \text{mm!} \]

The synchronization system is more than adequate.

Figure 17
Laboratory, under contract to the M.I.T. Magneplane group, and from a Ford Motor report, under contract to the Federal Railroad Administration (6,7) (see Table 1). Using root-locus techniques, similar to those used in the design of the model vehicle's control system, I arrived at the model and the calculations shown in figure 18. The model is a third-order system with feedback of velocity and position. This system is suitable for the acceleration and deceleration modes. It is critically damped, with poles with real parts at 5, 3 and $3 \text{sec}^{-1}$. The velocity error during acceleration does lead to constantly increasing position error, but this is acceptable, because there is independent synchronization control during acceleration.

The motor model ignores the motor force's dependence on position, because the propulsion winding field is expected to be predominantly the sum of a fundamental and a third harmonic, with a moderately flat peak amplitude from $-60^\circ$ to $+60^\circ$ around the peak of the fundamental. When the system leaves this region during cruise, the system will be in danger of loss of synchronization. However, once the vehicle slips back a pole pitch, it will not necessarily lose synchronization for good, as a conventional machine would. Because there is information feedback, the system can shut off current to the track while the vehicle slips $180^\circ$, then restore current during the entire $180^\circ$ of positive force, until the vehicle returns to a velocity corresponding to 60 Hz. Synchronization will not be irretrievably lost until the slip frequency of the vehicle with respect to the armature field is comparable to the inverse of the characteristic delay time of the control system.

For the cruising region, the system should include feedback of the
### Table 1

**Full Scale Vehicle Parameters**

**Vehicle**

- Mass - 40,000 Kg
- No. of alternating magnets - 24
- Maximum flux density at guideway - .25 w/m²
- Pole pitch - 1m
- Lift-to-drag - 40

**Single Section of Propulsion Winding, per phase**

- Resistance - .226Ω
- Inductance - 2.13 mh
- Pole Pitch - 1m
- Propulsion section length - 1Km
- Conductor width - .25m

**System Characteristics**

- Vehicle velocity - 120m/sec.
- Aerodynamic drag power loss - 3Mw
- $I^2R$ losses - .737Mw
- Magnetic drag power loss - 1.178Mw
- Total real power - 4.915Mw
- RMS phase Current - 3300A
- Power frequency - 60 Hz.
Acceleration Control with Velocity and Position Feedback

\[ V_{\text{REF}} \]

\[ \Sigma \]

\[ K_v + \frac{k_p}{s} \]

\[ E_a \]

\[ \frac{1}{sL + R} \]

\[ I \]

\[ K \]

\[ \frac{1}{sM + G} \]

\[ \text{VELOCITY} \]

\[ K_v, k_p \] = compensation gains for velocity, position feedback

\[ L, R \] = per phase inductance, resistance of the armature

\[ K \] = the motor constant, (m/sec)/newton, volts/(m/sec)

\[ M \] = mass of the vehicle

\[ G \] = incremental drag on the vehicle, newtons/m/sec

\[ V_{\text{REF}} \] = desired velocity

\[
\frac{V}{V_{\text{REF}}} = \frac{\beta s}{s^3 + (w^2_{e} + w^2_{m}) s^2 + (K_v w_{e} + K_v w_{m}) s + k \beta}
\]

where \( \beta = K/LM = .149 \text{ sec}^{-2} \)

\[ w_{e} = R/L, \ w_{m} = G/M \]

Let \( k_p = 300, \ K_v = 100 \)

\[
\frac{V}{V_{\text{REF}}} = \frac{.149 s}{s^3 + 10.6 s^2 + 37.4 s + 44.8} = \frac{.149 s}{(s+4.5)(s^2+6.1+10)}
\]

\[ \zeta = 1 \quad \omega_n = 3.1 \]

The system is critically damped. Response time is two orders of magnitude less than velocity decay time constant.

Figure 18
integral of the position, so that during steady-state cruise there will be no position error and the motor will be farther inside the stable synchronization region at all times. A third-order model was created by neglecting the short L/R time constant of the propulsion windings. The best system so far obtained, (see figure 19) is critically damped, but has a characteristic frequency of only 1 per 10 seconds.

Many of the serious problems concerned with the control system can be examined by considering the maintenance of synchronization in the face of strong head winds at cruising. Using the Raytheon parameters and assuming that aerodynamic drag is proportional to the square of the relative velocity of the vehicle, a sudden 50 mph wind would take 2 seconds to retard the motion of the vehicle by 1/2 meter. A 100 mph wind gust would take 1.4 seconds. Therefore, in order to prevent the vehicle from lagging by as much as 1/2 meter, or 90°, the dominant time constant of the control system should be no greater than a second or two. The system suggested in figure has two poles at $f=1/10$ second, or $\tau = 1.5$ seconds. Preventing the position error from ever exceeding 1/2 a pole pitch is an unnecessarily severe constraint, and the first simple controls looked at could be expected to achieve it, if marginally.

The additional current necessary to maintain synchronous velocity does not appear to be a major problem. If a 50 mph headwind is sustained, the necessary steady-state consumption of current would be 4200 A rms, up 27% from no head wind current. For a 100 mph headwind, 5200 A would be required, up 58%. The aluminum in the propulsion winding is derated in respect to current-carrying capability, because it is also used for
Velocity Control with Velocity, Position and Integral Position Feedback

\[ V_{\text{ref}} \]

\[ V_{\text{ref}} + \frac{K_V + k_p \frac{K_p}{s} + k_{pi} \frac{K_p}{s^2}}{s} \]

\[ V_e \]

\[ \frac{I}{R} \]

\[ K \]

\[ \frac{1}{SM+G} \]

\[ \alpha \]

\[ \frac{s + \omega_m + \alpha K}{s + \omega_m + \alpha K} \]

\[ V_{\text{ref}} = \text{desired velocity} \]

\[ K_V, k_p, k_{pi} = \text{velocity, position and integral position compensation gain constants} \]

\[ R = \text{per phase resistance of the armature winding} \]

\[ K = \text{the motor constant, Newtons/Amp} \]

\[ M = \text{the vehicle mass} \]

\[ G = \text{the vehicle incremental drag, Newtons/m.sec}^{-1} \]

\[ \alpha = \frac{K}{RC} = 0.014 \text{m.sec}^{-1}/\text{volt} \]

\[ \omega_m = 0.0084 \text{ sec}^{-1}, \omega_m \text{ can be neglected} \]

\[ \frac{v}{V_{\text{ref}}} = \frac{G}{1+G} = \frac{\alpha(Kv + k_p s^2 + k_{pi} s)}{s^3 + (\alpha K + K_v) s^2 + \alpha_k s + \alpha_{pi}} \]

Let \( K_V = 300, k_p = 300, k_{pi} = 100 = \frac{4.2(s^2 + s + 0.33)}{(s+3.2)(s^2 + 1.2s + 0.437)} \)

\( \zeta = 0.91, \quad \omega_n = 0.66 \)

Figure 19
structural support. Even if the motor were running hot, and could not run hotter during headwinds, the cost of propulsion winding aluminum is only 20% of the total system aluminum cost. If it had to be increased 25% to handle increased current during headwinds, the total aluminum cost would increase by 5%, and the total system cost by as little as 1 or 2%.

The power cost of control does not appear to be an important factor. The high efficiency of the motor ensures that total voltage will not rise greatly under any circumstances. Since the motor model has an efficiency of 85%, if current should double during headwinds, total per phase voltage would increase only 15%. Similarly, the federal ride standard requirement that longitudinal jerk be less than .03g means that, under constant drag, current rate-of-change $di/dt$ must be less than 1000 A/sec. This corresponds to an insignificant L $di/dt$ voltage rise of 2 volts.

While the voltage ratings of power conditioning units do not have to be significantly derated for overloads, current surges will require a little more attention. Federal ride standards allow .1g of sustained acceleration. However, at cruising, if the vehicle should lag, $1 m/sec^2$ of restoring acceleration will double the current in the track. If the vehicle were periodically accelerating and decelerating at this rate, the r.m.s. current would increase by $\left(\frac{[I+\Delta I]^2/2+(I-\Delta I)^2/2]^{1/2}-I}{I}\right)^2$ For $\Delta I=I$, this would equal an increase of 40%. If the vehicle lags by 60° or 1/3 meter, and the control system is designed to restore it to the maximum thrust position in 2 seconds, an acceleration of $0.2 m/sec^2$ is necessary. This would require only a 2.5% increase in the rms current.
Thus the power conditioning units would not have to be derated much more than the 10-20% derating necessary to combat head winds. Since two seconds is faster than the vehicle response time of the suggested control system (figure 19), it might be reasonable to limit control acceleration at cruising to less than federal standards.

Dr. Henry Kolm (9) has suggested that a full-scale system might be run in partially evacuated pipes. This would greatly ease the longitudinal motion control requirements by decreasing total aerodynamic drag and greatly lowering the possibility of unexpected cross and head winds. However, by decreasing drag, and therefore current, possibly by a factor of five, the control of heave motion by trimming the phase of armature currents would become much more difficult, if not impossible. Analysis of active heave damping has been ignored because of its complexity, and must be considered in another study.
Conclusions

Phase-locked loop motor control has been demonstrated to be a valid method for accelerating a linear synchronous motor. The limitations of the concept have not been determined, because of experimental difficulties. When all the fabrication errors have been eliminated, it is expected that velocity will be limited principally by the current limit of the cycloconverter. Strip chart recordings will then be used to determine whether the present system senses and controls the relative phase of the vehicle and armature phases with much precision. Recordings of the actuating signal (Butterworth output) made at Wayland on the samarium-cobalt vehicle show that thyristor transients and magnetic field "jumping" due to faithful reproduction of the square wave control input remain the primary source of noise in the system. At present, they are an order of magnitude greater in r.m.s. amplitude than the desired "average over a few cycles" error signal. The input to the 8038 has instantaneous voltages up ± 50% of the average voltage at a limiting velocity between 15-20 m.p.h.

In order to increase the vehicle velocity to 45 m.p.h., achieve levitation and damp vertical oscillations, the design may have to be improved so that the theoretical limits of phase-locked loop control are more nearly attained. Present plans are to optimize system open-loop gains and to add peak-clipping to the phase sensor to further eliminate transients without adding phase shift.

A full-scale system should employ other sensors, because reliability requires redundancy. In particular, a magnetic phase sensing device cannot realize when the vehicle has slipped a cycle, nor can it
reliably start smoothly. However, it is likely that a full-scale system will be run as a phase-locked loop with gaussmeters detecting vehicle and armature relative phase. The control equations presented indicate that such phase-locked loop control can achieve rapid control of synchronization by being able to control the thrust angle by "moving" currents instead of masses. The principle has been demonstrated successfully on a scale model, and will be perfected to include stable, levitated cruising.
APPENDIX I

The Cycloconverter

The cycloconverter and its attendant logic circuitry were assembled by Raytheon, using Westamp servo amplifiers. The circuit diagrams used in this chapter were drawn and conceived by Sumner Brown. The cycloconverter employs 18 SCRs, 6 for each input phase and 6 for each motor phase. The circuitry for each motor phase is mounted in one of three identical blocks, so that the entire cycloconverter can be understood by understanding one phase. (See figure 20) Each phase contains 3 pairs of SCRs. Each pair is connected between one phase of the 400 Hz input and the motor phase connected to the board. The two SCRs in a pair have different polarities, so that current can flow in both directions through the propulsion winding. Current is returned through the motor winding to a neutral, common to all 3 motor phases.

Each line phase on each board also contains a large series inductor to limit $di/dt$ through the SCRs, a shunt 2 µf capacitor and 47 ohm resistor to limit the $dv/dt$ across an SCR, a transformer tap to inform the main logic board of the polarity of the supply voltage, and a pulse transformer for triggering each SCR, with a free-wheeling diode to prevent gate breakdown by a high $L \frac{di}{dt}$ voltage when the pulse excitation is removed.

A conceptual diagram of the main board logic circuits was prepared by Sumner Brown (see figure 21), and it is to this diagram that I will refer. The transformer tap from phase A of the 400 cycle input feeds a low-level signal, proportional to the power input voltage into a rectifier and 2 comparators. The output of the rectifier gives a
WESTAMP CYCLOCONVERTER MAIN BOARD
\( \phi A \) FIRING CONTROL CONCEPTUAL DIAGRAM

CONTROL INPUT 1
[AMPLITUDE]

\[ \phi A \]

RECTIFIER
POLARITY 1
POLARITY 2

800 k

390 \( \Omega \)

T.P. 4, 7
10

100 \( \Omega \)

RESET

SET (FIRE)

F.F. OUTPUT

FLIP FLOP

HIGH PASS FILTER (DETERMINES TRIGGER PULSE LENGTH)

CONTROL [POLARITY] POSITIVE

INPUT 2 HIGH FOR OUTPUT

T.P. 13

AND

T.P. 5, 8, 11

1

\#1

\( \phi A \) (POWER IN)

\( \phi A \) POLARITY

RESULT

<table>
<thead>
<tr>
<th>CONTROL INPUT 2</th>
<th>( \phi A ) POLARITY</th>
<th>RESULT</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>+</td>
<td>FIRE 2</td>
</tr>
<tr>
<td>1</td>
<td>-</td>
<td>NOTHING</td>
</tr>
<tr>
<td>0</td>
<td>+</td>
<td>NOTHING</td>
</tr>
<tr>
<td>0</td>
<td>-</td>
<td>FIRE 1</td>
</tr>
</tbody>
</table>

FIGURE 21
signal whose instantaneous amplitude is directly proportional to that of the input phase voltage. The output of the comparator called "Polarity 1" is high when the line voltage is low. The comparator called "Polarity 2" is high when the line voltage is high. Control inputs 1 and 2 are formed by a signal conditioning circuit. A low-level signal which is proportional to the desired voltage across the motor phase is rectified and its current level is amplified by a compensated gain of one transistor. This creates control input 1, which is the desired instantaneous amplitude of the armature voltage. The unrectified signal is sent through a comparator whose output is high when the desired motor phase voltage is greater than zero. The amplitude levels and time constants are such that, if there is no control input 1, the output of the line phase rectifier will charge up the capacitor at such a rate that the unijunction will fire shortly before the phase A input has gone to 180°. Control input 1 is always positive, so it always advances the time during the half-cycle, when the unijunction fires. When control input 1 is at the maximum permissible amplitude, the unijunction fires almost immediately at the beginning of the half-cycle, waiting only for the flip-flop to reset. When the unijunction fires, it drives the output of a flip-flop high. The flip-flop drives the collector of a transistor to ground, discharging the timing capacitor and preparing it for a new charging cycle. The flip-flop output is also sent through a high-pass filter, whose time constant determines the duration of the trigger pulses to the SCRs. This pulse is then added with the outputs from control input 2 and Polarity 2 at the drive for SCR #2, the SCR which delivers positive voltage to the track. In other words, the AND gate asks: is the line-
voltage positive? Do we want positive motor voltage? Have we waited long enough that the average voltage across the armature will equal the desired voltage? When the answer to all three questions is yes, SCR2 is fired. Similarly, AND gate one asks about the reference timing and whether the desired and line voltage are both negative. The waveforms of a circuit whose reference voltage has a frequency equal to 1/4 of the line frequency are displayed in figures 22 and 23. A resistive load is assumed. As can be seen, even at a frequency as low as 1/4 of the supply frequency, the output waveform has lost all visual resemblance to the desired waveform, but the power spectrum of the output is still very close to that of the input.
CYCLOCONVERTER LOGIC CIRCUIT WAVEFORMS

ΦA

ΦB

ΦC

RECT. A

POL. 2

POL. 1

CONTROL
INPUT 1

YCAP

SET

FF OUT

FIGURE 22
Figure 23

Cycloconverter Logic Circuit Waveforms

- High Pass
- Control Input 2
- AND 1 OUT
- AND 2 OUT
- ω A OUT
- ω B OUT
- ω C OUT
- Vout
APPENDIX II

THE TRANSMITTER CIRCUIT

FIGURE 24
THE VEHICLE CIRCUIT

Figure 25
BIBLIOGRAPHY


