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AROD
SPECIAL TECHNICAL REPORT
VEHICLE TRACKING RECEIVER DESIGN

MOTOROLA INC.
Government Electronics Division
Aerospace Center

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AROD

VEHICLE TRACKING RECEIVER DESIGN

1.0 INTRODUCTION

The Vehicle Tracking Receiver incorporates several unusual applications of phase-locked loops and signal processing techniques. The purposes of this report are to document the design characteristics of the receiver, to present the constraints upon the design, and to provide a background for evaluation of the equipment performance. Included in this report are some detailed performance data in the form of curves, photographs, and recordings which were measured on a breadboard version of the receiver operating with a breadboard Transponder Transmitter. This data is presented in the section of this report detailing that portion of the receiver design.

1.1 DESCRIPTION OF RECEIVER

The receiver is comprised of four receiving channels which operate independently. A portion of the receiver, specifically the S-Band Mixer and Preamplifier, the Frequency Synthesizer, and assorted Distribution Amplifiers, is common to the four channels. Excluding the common equipment, the equipment in the remainder of the receiver channels is identical for each channel except for tuning and the selection of reference frequencies. The non-common portion of each receiver channel is physically constructed in three subsections. These are the Carrier Tracking Loop, the Modulation Tracking Loop, and the Code Control Subsection. These subsections operate together to provide amplification, demodulation, and tracking of the input signal. Initial acquisition and station release is controlled by the AROD System Control Logic which provides appropriate steering signals to the receiver.

The received signal is transmitted from a Transponder Station as a suppressed-carrier, pseudonoise-coded waveform. During initial acquisition, the transponder has applied Doppler compensation to the carrier frequency of this signal to ensure that the buried carrier lies within ready capture range of the Vehicle Receiver. The impressed code during this initial acquisition is a 127 bit code with a bit rate of 6.25 KHz. This code is searched at a rate of 50 bits per second in the receiver resulting in an inspection time of 20 milliseconds for each code bit.

The receiver is programmed to Acquisition State V-1 to perform this code search. The receiver requirements in this state are:

(1) Rapid capture of code and carrier,

(2) Rapid detection of code synchronization,

(3) Automatic termination of the code search, and

(4) Demodulation of the data on the signal to determine station identification and status.

After the initial code and carrier acquisition, the Doppler compensation is removed by the Transponder Station with a linear sweep from Reversed Doppler to True Doppler. The receiver tracks this sweep in Acquisition State V-2 while maintaining code lock. The receiver requirements in this state are:

(1) Track the frequency sweep with acceptable lag,

(2) Maintain code synchronization, and

(3) Demodulate the data on the signal indicating the completion of the sweep.

When the Doppler Reverse sweep is complete, the Transponder Station signal is modified to include a high frequency code. This code is 511 bits long with a bit rate of 6.4 MHz. This entire High code sequence is included in each bit of the Low code which was initially acquired. Upon recognition of the Doppler Reverse Complete signal, the Vehicle Receiver advances to Acquisition State V-3 for capture of the High code. In this state the receiver requirements are:

(1) Continue to track the suppressed carrier,

(2) Continue to track the Low code,

(3) Perform a restricted search for the High code, and

(4) Detect High code synchronization.

When High code synchronization is detected the receiver is automatically advanced to the final Track mode V-4. In this state the receiver tracks the suppressed carrier and the High code. The Low code is retained in the transmitted signal, but is not used for tracking in this mode. The receiver requirements in Acquisition State V-4 are:

(1) Track the suppressed carrier through the expected signal environment due to vehicle dynamics, signal fading, and Transponder to Vehicle propagation loss,

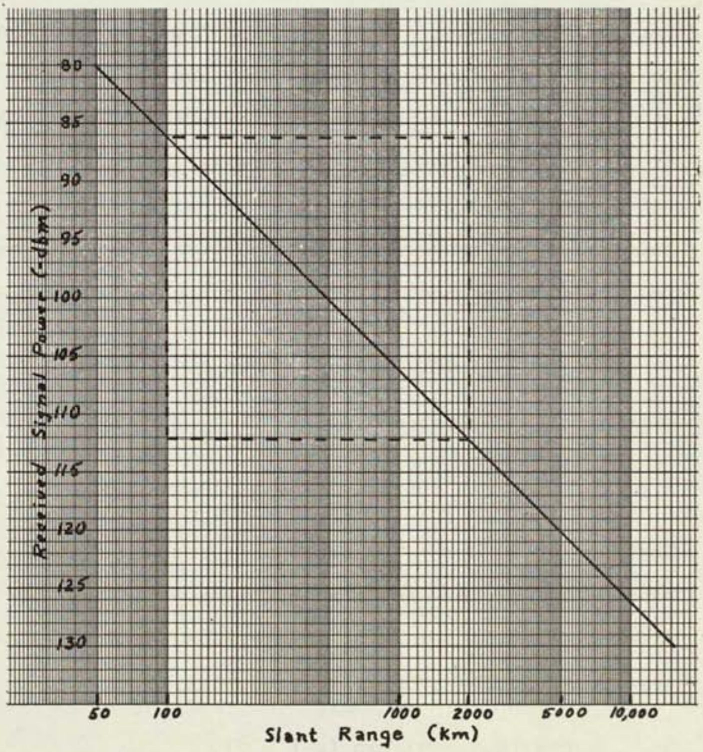
(2) Perform optimum filtering of the carrier and extract the carrier Doppler shift for determination of Transponder to Vehicle velocity data,

(3) Track the High code and perform optimum filtering of its phase for extraction of Transponder to Vehicle range data, and

(4) Detect data on the received signal indicating Transponder identification and status.

The received signal power at the Vehicle Tracking Receiver is shown graphically in Figure 1. The region of primary interest is enclosed in the dashed line box. This region covers slant ranges from 100 KM to 2000 KM. The antenna gains, transmitted power, and cable losses are those anticipated for the System Test Model and may not be representative of operational conditions.

Frequency	1800 MHz
Transmitter Power	+43 dbm
Transmitter Mismatch & Cable Loss (10' RG-9)	-1.4 db
Station Antenna Gain (minimum)	+11 db
Vehicle Antenna Gain	+3 db
Polarization Loss	-3 db
Receiver Mismatch & Cable Loss (5' RG-223)	-1.2 db



Vehicle Tracking Receiver
Signal Level vs Range

FIGURE 1

2.0 MODULATION PROCESSING AND CARRIER TRACKING

The Tracking Receiver is intended to track the carrier frequency of the incoming signal for coherent velocity extraction, and to recover the range code modulation for the extraction of range data. To accomplish these tasks, the receiver design is based upon the principle that the carrier and the modulation are best extracted through two parallel paths. This principle can be summarized by considering a received signal which is the product of a carrier and its modulation, or $M(t) \cos (\omega_c t + \theta)$ and the parallel extraction paths being the multiplication of this signal by two different references for range and velocity extraction. The reference for the extraction of the carrier for velocity data is

$$M(t + \tau) \frac{d}{dt} \cos (\omega_c t + \phi)$$

which yields the product

$$\left[M(t)\ M(t + \tau)\right] \left[\cos (\omega_c t + \theta) \sin (\omega_c t + \phi)\right]$$

The term $\left[M(t)\ M(t + \tau)\right]$ represents the correlation of the ranging signal and its cross-coupling into the velocity extraction. The remaining term yields the desired carrier tracking function, for by trigonometry,

$$\cos (\omega_c t + \theta) \sin (\omega_c t + \phi)$$

$$= \frac{1}{2} \left[\sin (2\omega_c t + \theta + \phi) + \sin (\phi - \theta)\right]$$

This is seen to be the sum of two terms, one of which is a high frequency signal at twice the carrier frequency, and the other related only to the phase error between the incoming carrier and the locally generated replica. These signals can be separated by a low pass filter which yields the desired error signal of

$$A \sin (\phi - \theta)$$

The carrier channel thus develops a steering signal which is proportional to the sine of the phase error. This steering signal is properly sign-sensitive (i.e., positive phase error leads to positive voltage and negative phase error leads to negative voltage) and reduces to zero when tracking with no phase error.

The modulation is recovered by a similar means. As the modulation may in general be represented as

$$M(t) = \sum_{n=0}^{N} a_n \cos (\omega_n t + \alpha_n)$$

the received signal is

$$S = \sum_{n=0}^{N} a_n \cos (\omega_n t + \alpha_n) \cos (\omega_c t + \theta)$$

The desired reference is

$$R = \frac{d}{dt} M(t + \tau) \cos (\omega_c t + \phi)$$

which can be represented by

$$R = \sum_{m=0}^{M} b_m \sin (\omega_m t + \beta_m) \cos (\omega_c t + \phi)$$

upon multiplication with the received signal, the result is

$$\cos (\omega_c t + \theta) \cos (\omega_c t + \phi) \sum_{0}^{N} \sum_{0}^{M} a_n b_m \cos (\omega_n t + \alpha_n) \sin (\omega_m t + \beta_m)$$

The average value of which is everywhere zero except when m = n. The low frequency part is then

$$\cos (\omega_c t + \theta) \cos (\omega_c t + \phi) \sum_{n=0}^{N} c_n \cos (\omega_n t + \alpha_n) \sin (\omega_n t + \beta_n)$$

Employing the trigonometric identities used in the derivation of the carrier steering signal, the term $\cos (\omega_c t + \theta) \cos (\omega_c t + \phi)$, after selection of the low frequency terms, yields $\cos (\theta - \phi)$ and represents the degree of correlation of the carrier signals and its cross-coupling into the modulation tracking system.

The remaining terms inside the summation are all of the form cos x sin y which is equal to

$$\frac{1}{2}\ \sin. (x + y) + \sin (x - y)$$

and is the sum of signals at twice the frequency of ω_n and phase error terms which can be selected by appropriate low pass filters.

The resultant low frequency part of the modulation steering signals is then

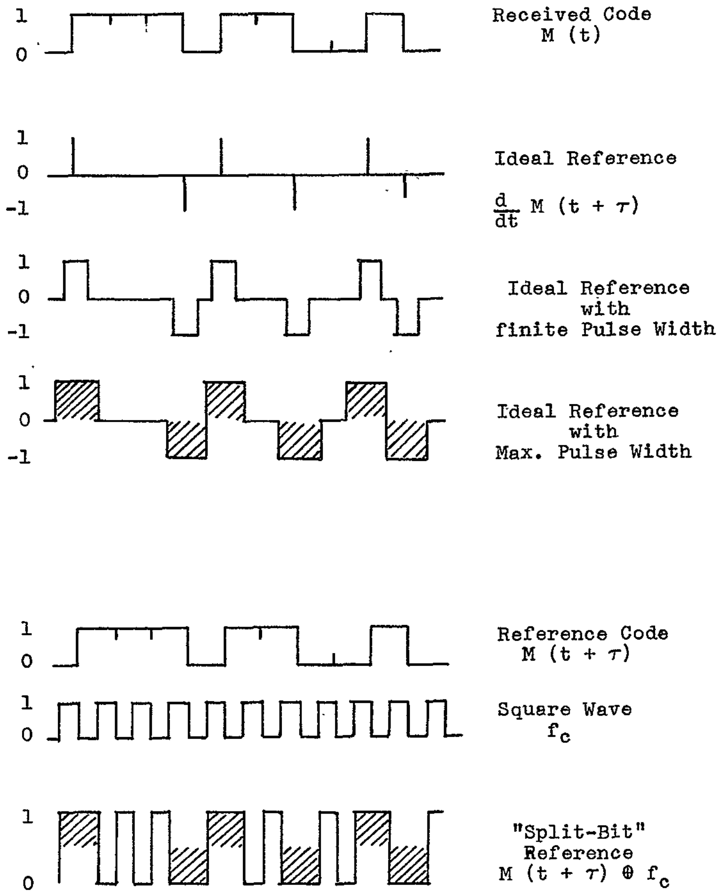
$$\frac{1}{2}\sum_{o}^{N} c_n \sin (\alpha - \beta)$$

which is proportional to the phase error between each component of the modulation signal and is properly sign sensitive.

The physical generation of the desired reference signals required some compromise of the forgoing development. The carrier tracking reference is straight-forward, for if the modulation signal could be generated in the transmitter it can be erected in the receiver also. The development of the derivative of the carrier signal represents merely a 90 degree phase shift of the carrier which introduces no particular problems.

The modulation reference is more difficult for complex waveforms. As the selected modulation is a binary pseudonoise sequence, its derivative is a series of positive and negative impulses. This derivative can be approximated by a "split-bit" whereby the impulses are stretched and balanced to ease the mechanization problems of three-level modulation and infinite bandwidth.

The relationship of a "split-bit" reference with an ideal reference is shown in Figure 2. There are two major differences. The first of these is the stretching of the impulses into a finite duration and the second is the substitution of a zero-mean square wave for the third level of the derivative reference. It should be noted that the frequency of the square wave, f_c, can be any integral multiple of the code clock frequency to establish the correspondence of the "split-bit" and derivative references. This characteristic is used in the design of the receiver.



A $\oplus$ B is defined as

B \ A	0	1
0	0	1
1	1	0

Relationship of "Split-Bit" to derivative Reference
FIGURE 2

2.1 SEQUENTIAL PROCESSING

The effect of multiplying the received signal by the split-bit reference does not necessarily affect the carrier tracking capability. Let the received signal be

$$AM(t) \cos (\omega_i t + \phi)$$

and the reference be

$$2M(t + \tau) \sqsubset\!\mathrm{os}\ \omega_c (t + \tau)$$

where

$$\sqsubset\!\mathrm{os}\ \omega_c (t + \tau)$$

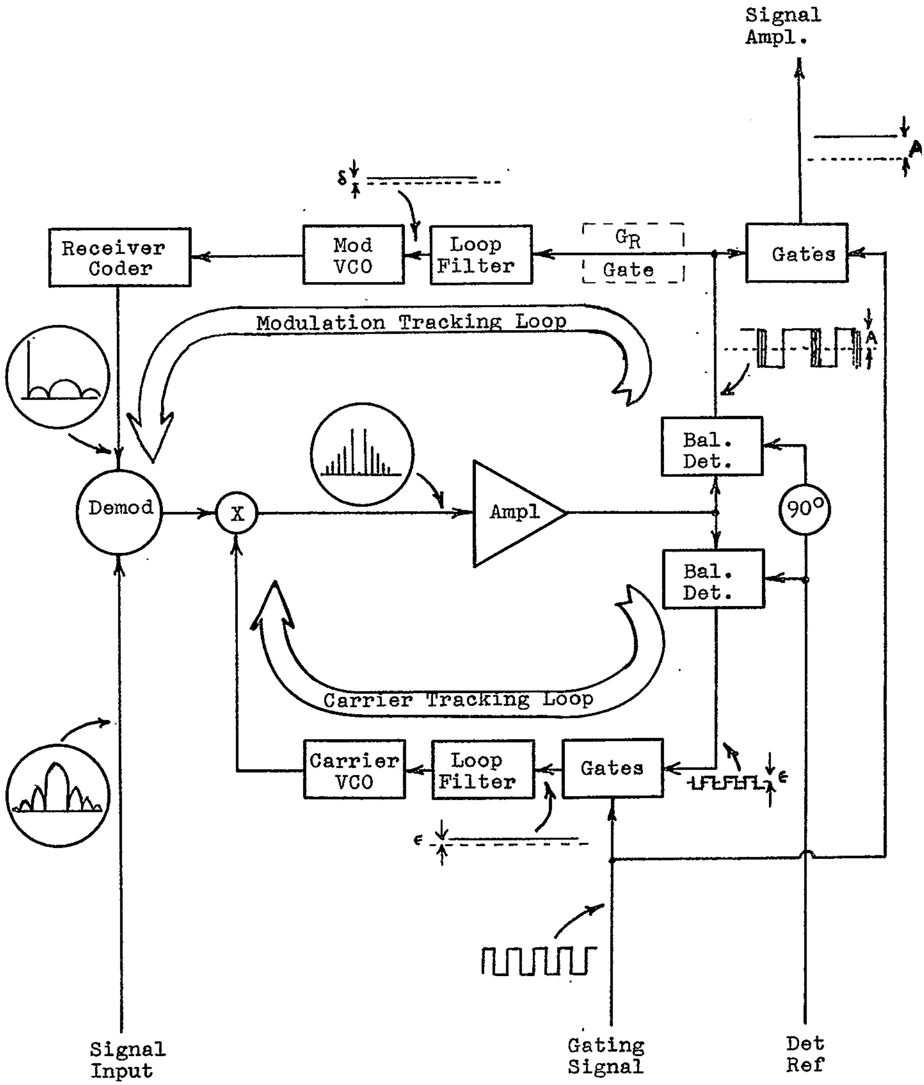
is a square wave with the timing of a $\cos \omega_c (t + \tau)$ function. The resultant product

$$AM(t)\, M (t + \tau) \sqsubset\!\mathrm{os}\ \omega_c (t + \tau) \cos (\omega_i t + \phi)$$

can be amplified, translated, and gain leveled before separation into the modulation and carrier error detectors. The low frequency terms after multiplication by $\cos (\omega_i t + \theta)$ was shown to be a proper modulation tracking signal, and the low frequency term after multiplication by $\sin (\omega_i t + \theta) \sqsubset\!\mathrm{os}\ \omega_c (t + \tau)$ yields the desired carrier error signal.

It is significant that the product of the received signal with the reference code contains timing information for the carrier. The spectrum of this signal is a sinusoid balance-modulated by the square wave. The residual center frequency amplitude is proportional to the modulation tracking error and its phase reverses when the modulation timing goes from early to late. In this form, the error signal for the modulation timing is highly insensitive to equipment drifts.

This method of processing the signal is shown in Figure 3. Also shown on this diagram is the method of recovering a measure of the signal amplitude. This amplitude is used to drive the automatic gain control circuitry and to provide a means of detecting code synchronization.



Basic Receiver Configuration

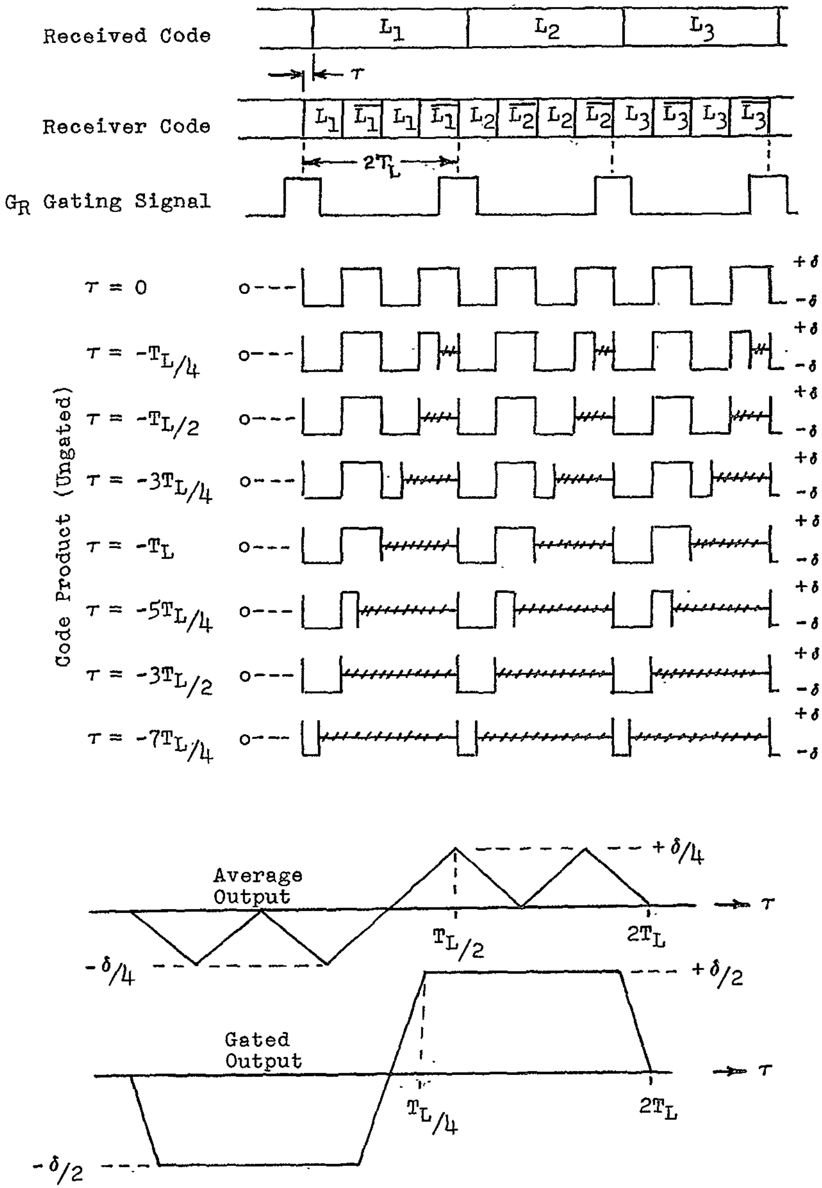
Figure 3

2.2 SELECTION OF GATING SIGNALS

The necessity of rapid automatic acquisition and tracking of both the low-speed code and the high-speed code requires a modification of the simple square wave gating signal which was shown to modulate the receiver code and to extract the error signals. The received code actually takes one of two forms, depending upon the particular portion of the acquisition sequence in process.

During initial acquisition, Vehicle States V-1 and V-2, the received code is simply the low code (L-code). The receiver reference code is modulated with a square wave which is twice the code bit rate. This requires that the modulation error signal be gated to remove the ambiguous nulls. The development of the modulation error signal as a function of the timing error is shown in Figure 4. The timing gate G_R is placed at the input to the modulation loop filter shown in Figure 3.

The L-code is acquired by the receiver by a sequential code search which causes the receiver code to shift relative to the received code one bit at a time. Code synchronization is detected by the presence of a signal at the output of the signal amplitude balanced gates. The carrier tracking loop during this code search is held to center frequency by a "Frequency Preset" mechanism and automatically locks when the correct code timing is achieved. When the L-code lock is complete and is so reported by the A_L detector, the S-Band



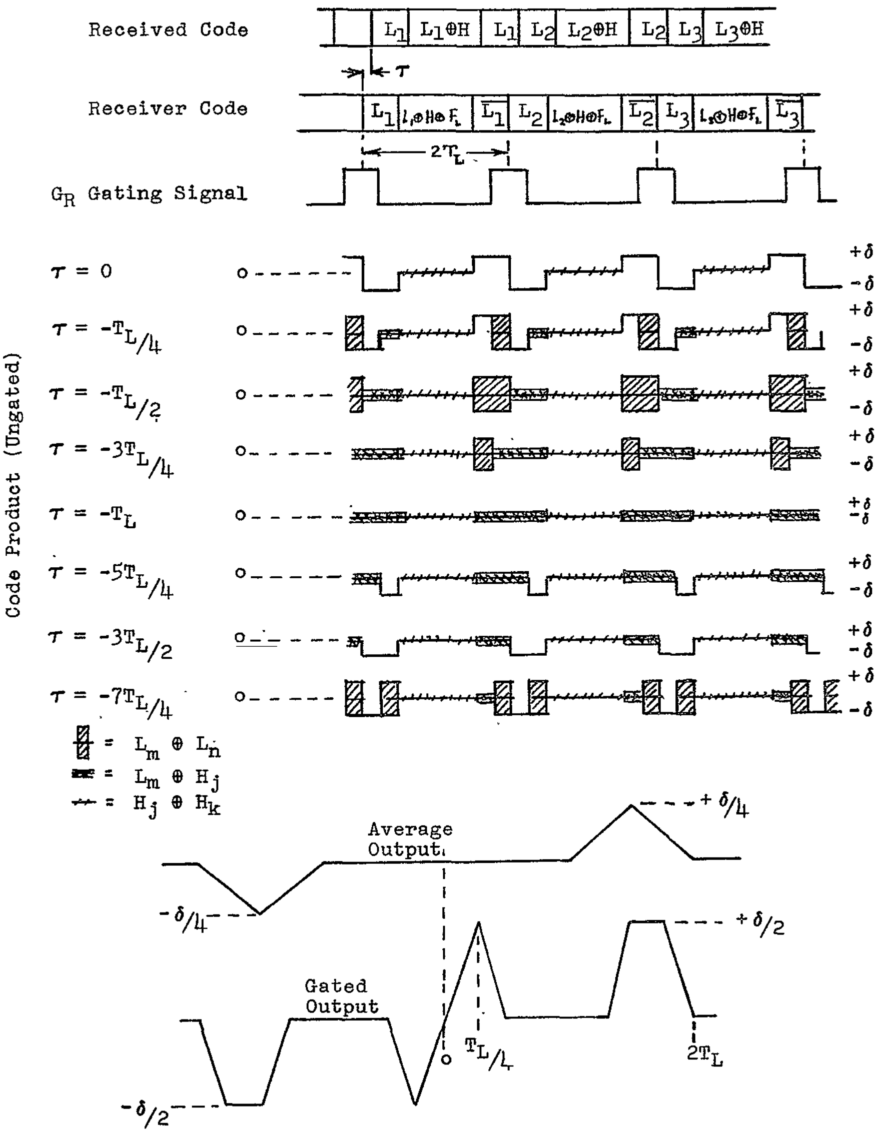
Modulation Tracking Signal
Acquisition State V-1 & V-2

Figure 4

frequency is linearly swept by the Transponder Station to establish two-way Doppler coherence. The Vehicle Receiver tracks this sweep in Acquisition State V-2. At the end of the sweep the Transponder Station modifies the transmitted code and simultaneously informs the receiver that the Doppler Reverse is complete. The Vehicle Receiver then advances to Acquisition State V-3 to acquire high-speed code. The development of the modulation tracking signal in State V-3 is shown in Figure 5. It can be seen that the modulation gating signal is necessary to generate a proper tracking signal in this mode. The presence of the residual "rabbit ears" is of no concern because the receiver timing has been correctly preset during the initial Acquisition State V-1.

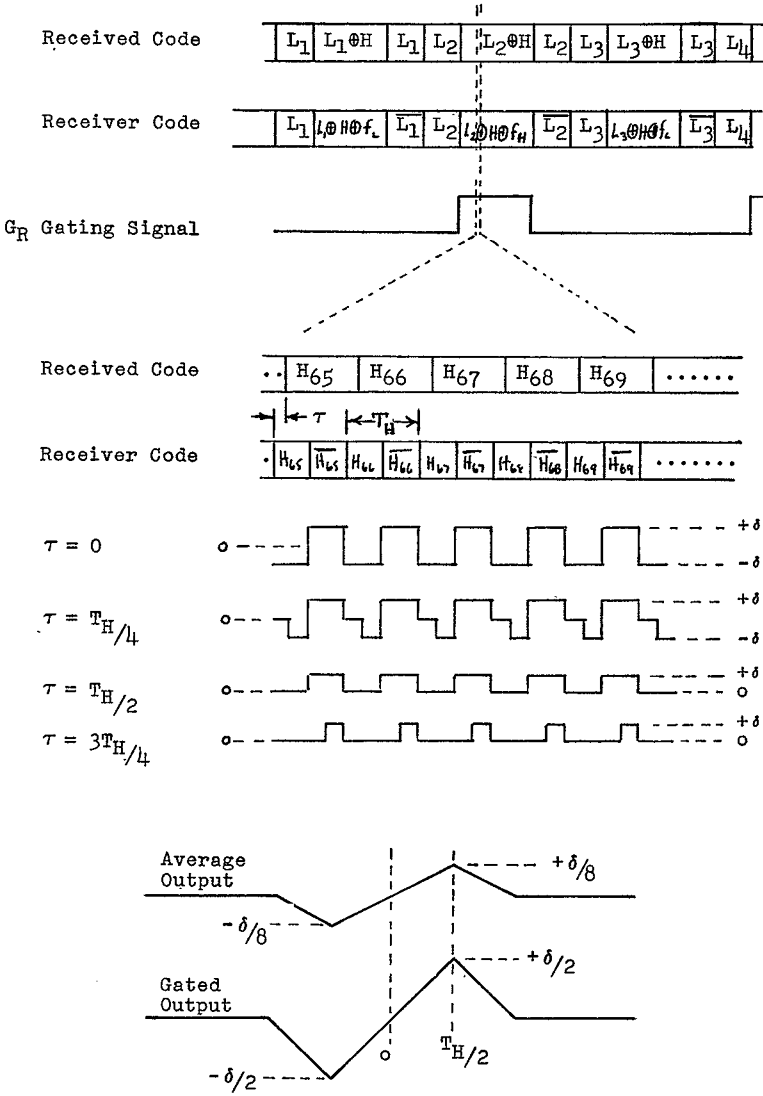
When the receiver is tracking the L-code in Acquisition State V-3, the signal amplitude is monitored both during the tracking period and in the interval of time where the H-codes are present. When H-code alignment is sensed in the A_H detector the receiver code and gating signals are switched to track on the H-code alone. The development of the modulation tracking signal in this final "Track" State V-4 is shown in Figure 6.

The application of these signals to the tracking receiver in the process of signal acquisition is shown in Figure 7. Included in this timing diagram is the effect of the i.f. gating signal to prohibit the L-code from introducing tracking errors in the final Track mode V-4.



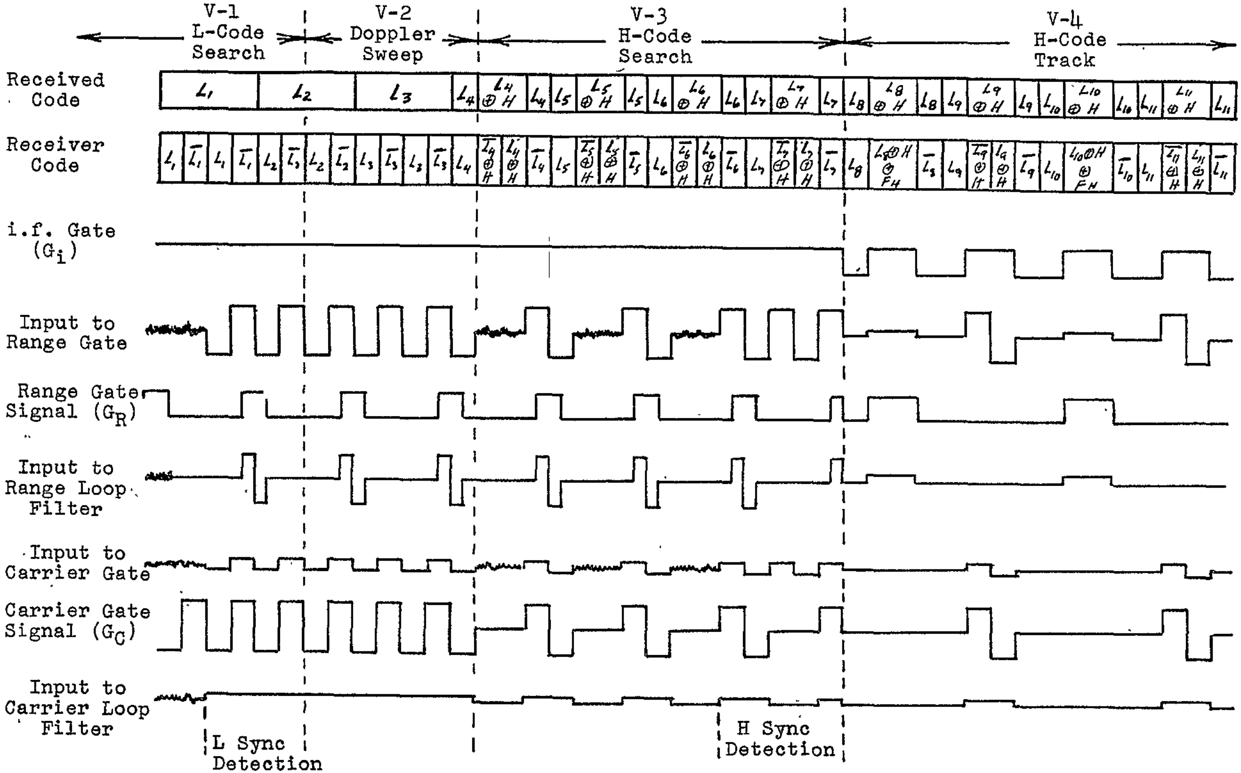
Modulation Tracking Signal
Acquisition State V-3

Figure 5



Modulation Tracking Signal
Acquisition State V-4

Figure 6



Receiver Signal Gating

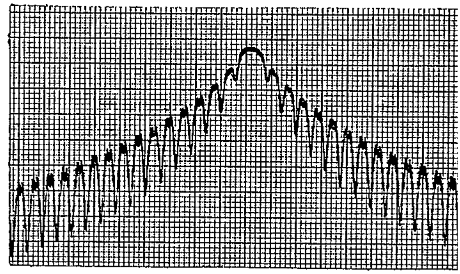
Figure 7

The tracking receiver in the final Track State V-4 is seen to be time shared between carrier tracking intervals, modulation tracking intervals, and intervals of time during which the L-code is received but the intermediate frequency amplifying chain is cut-off. One-half of the time is assigned to this normally unused L-code transmission to provide a means of rapid reacquisition in the event of signal discontinuities which cause loss of H-code lock. This protection affords the capability of continuing L-code lock by reverting to State V-3 without modifying the Transponder Station transmitted signal. The extent of this protection is apparent by considering that a timing error of ±.159 μsec ($\pm T_H$) causes loss of H-code lock, but the error must be increased to ±160 μsec ($\pm 2T_L$) to cause L-code loss.

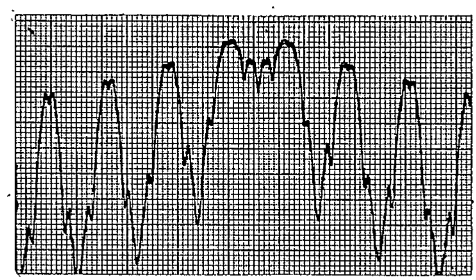
The other half of the time is equally shared between carrier track and modulation track intervals. These intervals both correspond to periods when the H-code is received from the Transponder Station. The receiver reference code is different for modulation tracking and carrier tracking in State V-4, being $L \oplus H \oplus f_H$ for modulation track and $L \oplus H \oplus f_L$ for carrier track. This processing differs from that used in the earlier acquisition states where one reference provided both tracking signals. The reason for the difference is that although the $L \oplus H \oplus f_H$ reference provides a proper split-bit reference and both carrier and modulation error signals result, the carrier tracking signal exists as the phase of sidebands about the normally suppressed center frequency and the sideband frequency is $\pm 2F_H$ or ± 12.8 MHz. It is not practical to support this bandwidth through the intermediate frequency channel as the noise entering the detectors would become excessive. Accordingly, the reference signal of $L \oplus H \oplus f_L$ is substituted for carrier tracking and the intermediate frequency sidebands are at $\pm f_L$ or ± 12.5 KHz which is the same frequency as the resultant in the earlier acquisition states.

The measured spectra of the received code, the reference code, and the resultant products are shown in Figures 8, 9, and 10. Although the time sharing in acquisition states V-3 and V-4 is apparent in the spectra as additional spectral lines, the continuity of tracking signals can be seen in the 12.5 KHz sidebands which are retained throughout the acquisition and tracking modes.

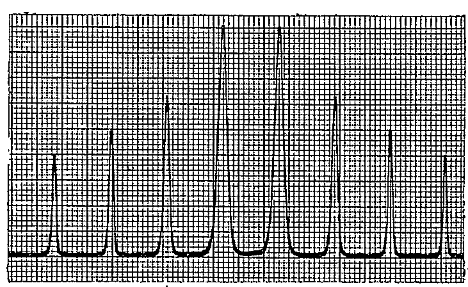
Receiver
Input
Spectrum



Receiver
Reference
Spectrum



Correlated
i.f. Spectrum



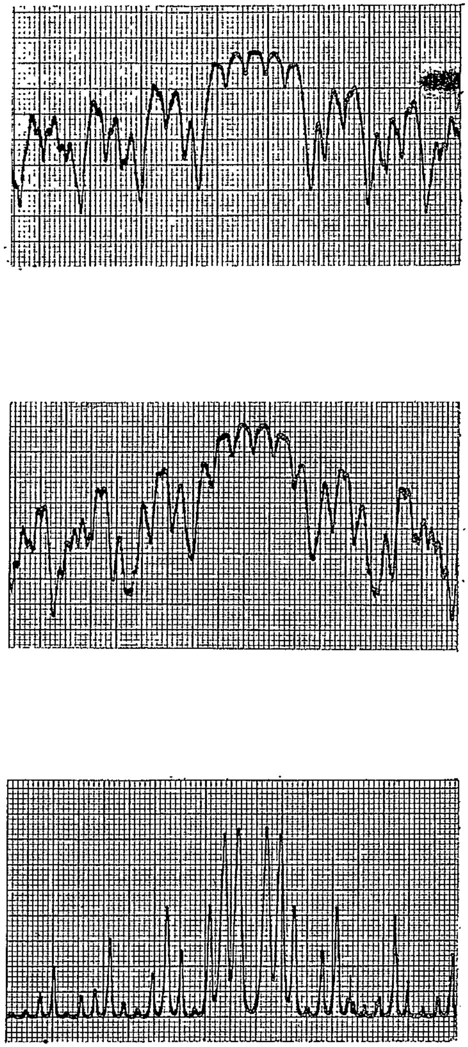
Horizontal: 55 KHz/in
Vertical: Logarithmic

Acquisition State V-2
Receiver Spectra
Figure 8

Receiver
Input
Spectrum

Receiver
Reference
Spectrum

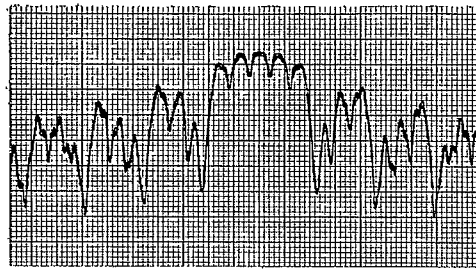
Correlated
i.f. Spectrum



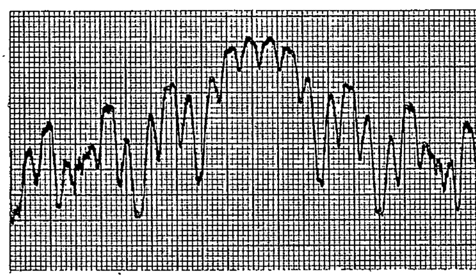
Horizontal: 55 KHz/in
Vertical: Logarithmic

Acquisition State V-3
Receiver Spectra
Figure 9

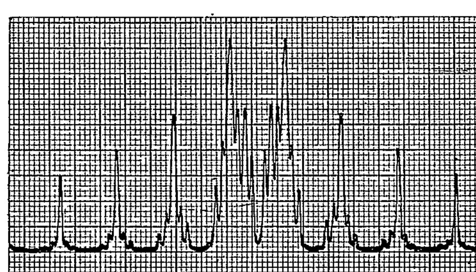
Receiver
Input
Spectrum



Receiver
Reference
Spectrum



Correlated
i.f. Spectrum



Horizontal: 55 KHz/in
Vertical: Logarithmic

Acquisition State V-4
Receiver Spectra
Figure 10

3.0 LOOP DESIGN

The Vehicle Tracking Receiver employs phase-locked loops for tracking of the received carrier and range modulation. The amplifying channel of the receiver is gain leveled by a coherent automatic gain control system, and a phase-locked Rate-Aid loop is used to assist the modulation loop in following signal dynamics. The following paragraphs are concerned with the constraints upon the design of these tracking loops.

3.1 LEVEL CONTROL AND AGC LOOP

The receiver is intended to operate over the range of input signal levels from -70 dbm to -130 dbm. Reference to Figure 1 shows that this is a much larger dynamic range than that expected for slant ranges of 100 km to 2000 km which is the primary region of interest. The larger dynamic range provides increased system versatility and protection against signal fading in operation. The receiver gain and amplifier distribution have been designed to afford this increased dynamic range and to control the magnitude of cross-channel intermodulation in the course of receiving multiple station responses.

The selection of the receiver intermediate frequencies, the order of mixing operations, and the gain control system are based upon the desired dynamic range, the relative power difference between station responses, the minimization of internally generated bias effects, and saturation due to thermal noise in the receiver.

The primary purpose of the AGC system in the receiver is to establish a relatively constant signal level into the phase detector driving the carrier phase-locked loop. This is necessary to control the carrier loop bandwidth between the limits required to track vehicle dynamics and to prevent carrier tracking of the range code modulation. The configuration of the amplifiers, detectors, and AGC system is shown in Figure 11.

The action of the automatic gain control system is to hold the level at the detector input nearly constant, however, the amplifier which drives the detectors will modify this level near receiver threshold due to the controlled limiting characteristics. This provides a means of allowing a predictable amount of loop bandwidth expansion in the receiver loops under normal conditions. The amplifier-limiter is a band limited circuit, and will hold nearly identical signal-to-noise ratios at the input and output, and as the total output power capability is restricted, the available signal power at the output will be decreased as the signal-to-noise ratio decreases. The automatic gain control system clamps the signal level very near the minimum value to which it is suppressed by the noise as shown in Figure 12a.

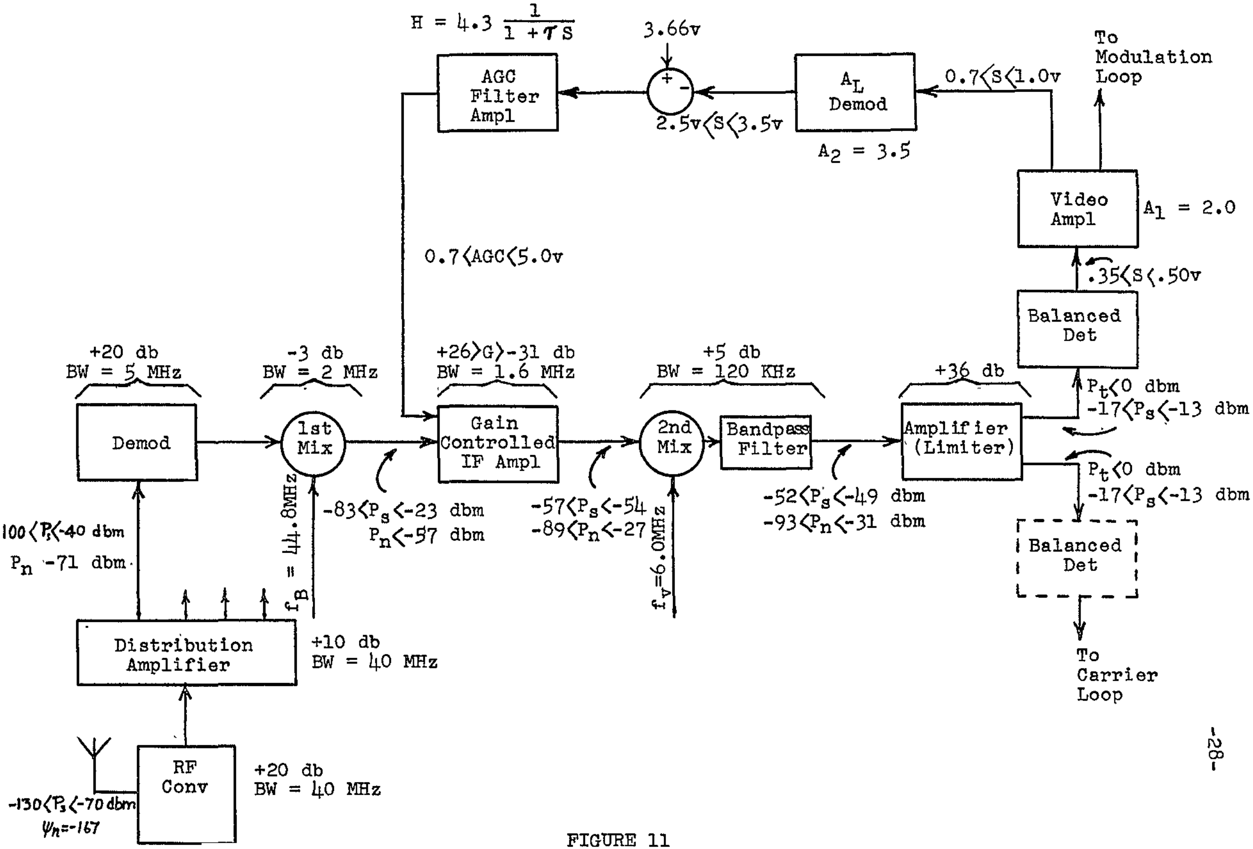
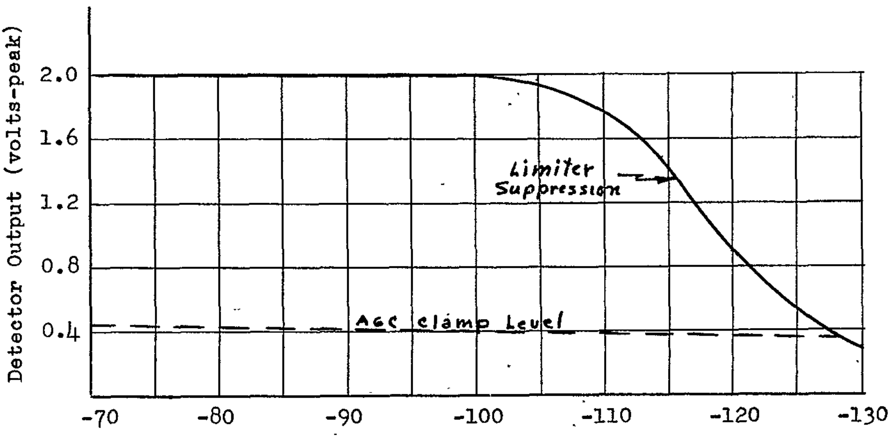
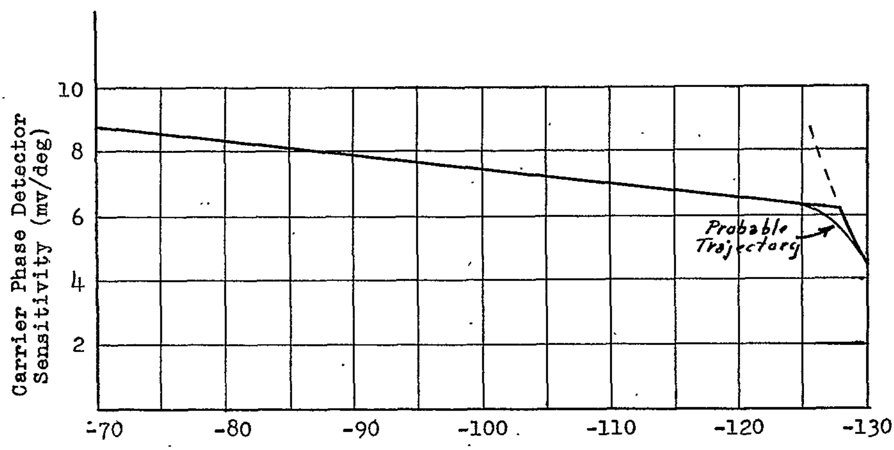


FIGURE 11
Receiver Gain Control



A

Predetection $BW_{3\ db}$ = 120 KHz
Noise Figure = 7.0 db



B

FIGURE 12

Combined Limiter and AGC Effects

The AGC loop is designed to be a first order servo with a transfer function of

$$H_{AGC} = \frac{1}{1 + KS}$$

where

$$K = \tau/G = \frac{\tau}{K_a K_d A}$$

$15 < K_a < 180$ db/v = gain slope of IF amplifier

$.037 < K_d < .05$ v/db = detector sensitivity

$A = 4.3 \times 3.5 \times 2.0$ = amplifier and gate gains

$\tau = 0.35$ sec = filter amplifier lag

The resultant noise bandwidth of the AGC loop will vary from 12 Hz at the receiver threshold to 160 Hz under very strong signal conditions.

Although the phase-locked carrier loop has been tightly constrained after acquisition, and its bandwidth has been restricted to prohibit code tracking, the coherent nature of this AGC offers no assistance during initial acquisition. To restrict the bandwidth under strong signal L-code search, a non-coherent detector supplies the AGC function. Under weak signal conditions and after acquisition, this detector is non-operative.

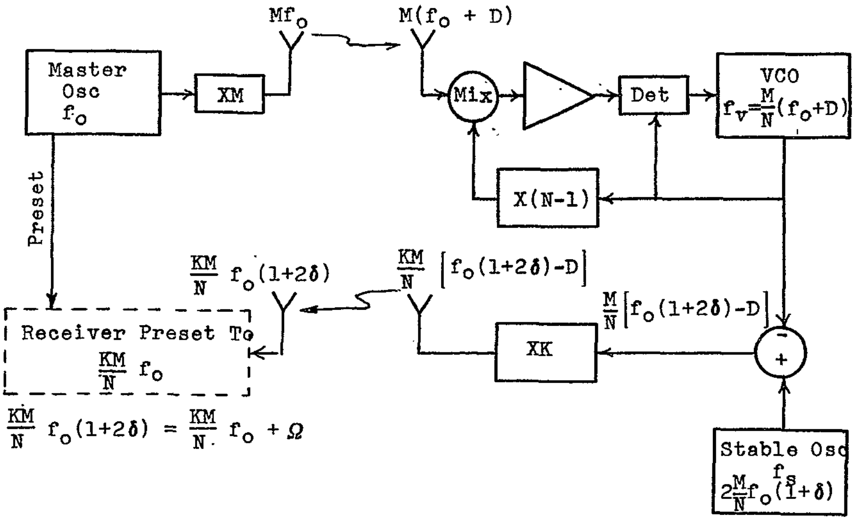
3.2 CARRIER LOOP DESIGN

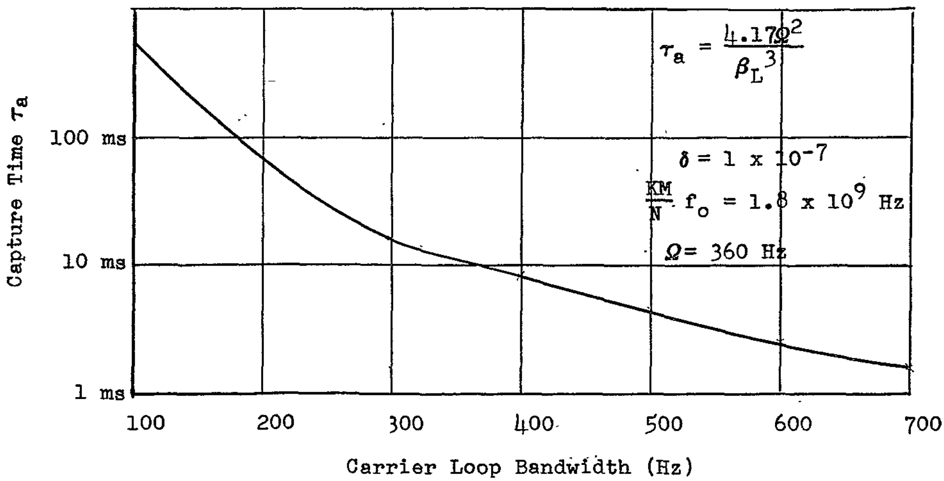
The design of the carrier tracking loop is constrained by the range of input frequencies to be tracked, the input signal dynamics, the time required for capture of the received signal, and the effects of the loop performance upon the code synchronization detectors. The resultant design is the result of a number of trade-offs between these constraints.

3.2.1 Bandwidth Selection

During initial acquisition the tracking receiver is to acquire the received signal in the Doppler Reversed mode. The L-code is searched at a rate of 50 bits/second and the carrier loop must acquire the signal upon the correct code bit. Upon capture, the coherent signal strength is detected in the A_L detector terminating the code search.

The Doppler Reversal is not exact, as there is no perfect measure of the vehicle master oscillator at the Transponder Station, so the received signal will differ from the receiver center frequency by an error. The receiver carrier VCO is programmed to center frequency by the Frequency Preset Circuitry with zero static error, so the frequency error can be attributed entirely to the accuracy of the Doppler Reversal. The equivalent model of the Doppler Reverse is shown in Figure 13, and the resultant capture time of the carrier loop is plotted for several carrier loop bandwidths.





Effect of Doppler Reverse Accuracy
Upon Carrier Bandwidth Selection

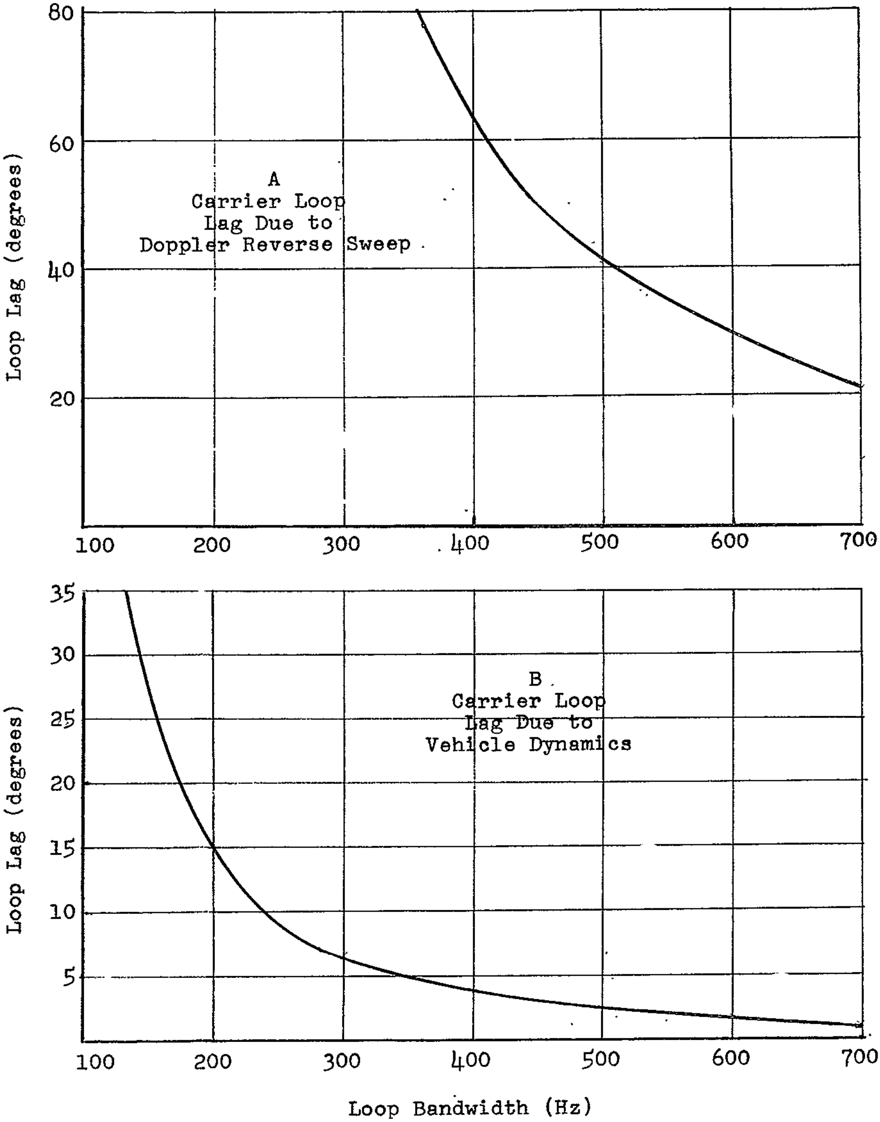
FIGURE 13

As there is only 20 ms available for inspection of each search step, it is seen that the loop capture time will degrade the acquisition threshold. The relationship is linear for the acquisition detector (A_L detector) is a matched filter with essentially linear characteristics. Acquisition bandwidths of the carrier loop must be greater than about 500 Hz to avoid a serious loss of acquisition sensitivity.

After initial acquisition, the Doppler Reversed signal is linearly swept to the true Doppler condition. This ramp of frequency is to be tracked by the vehicle receiver. The magnitude of the ramp may be estimated at escape velocity from the two-way Doppler shift of the 1800 MHz signal to be about 140 KHz, and the transponder station performs this sweep with a linear rate of 100 KHz/sec. The corresponding lag in the vehicle receiver will be a function of the carrier loop bandwidth as shown in Figure 14a.

When the Doppler Reverse cycle has been completed, the signal dynamics will be only those associated with the range-rate from the vehicle to the transponding station. This rate may be as high as 6 KHz/sec but will always occur under strong signal conditions.

The receiver lag to the Doppler rates from these vehicle dynamics is plotted as a function of receiver bandwidth in Figure 14b. It can be seen that a considerably narrower loop can be tolerated after acquisition has been



Effect of Signal Dynamics
on Carrier Loop Bandwidth Selection

FIGURE 14

completed.

There is an upper limit to the allowable carrier bandwidth which is a result of the type of modulation employed. As the code structure includes components down to about 50 Hz, and as the code is modulated directly upon the received carrier, the carrier loop will attempt to track a portion of the modulation. The wider the loop bandwidth, the more modulation is tracked, and the greater the degree of apparent code correlation even though the transmitter and receiver codes are not in synchronism. This complicates the problem of correct detection of code synchronization for the apparent code-noise level can become high under strong signal code search. The effects of this code tracking are minimized in the receiver by the three separate mechanisms.

(1) The basic bandwidth is minimized.

(2) The bandwidth expansion under strong signal conditions is controlled by non-coherent AGC.

(3) The carrier loop is constrained by the frequency preset.

The measured code correlation characteristics of the receiver for two basic carrier bandwidths, with and without the bandwidth expansion due to strong input signals, is shown in Figure 15. The bandwidth expansion can be calculated from

β_o = 1000 Hz

G/G_o = 8

β_o = 1000 Hz

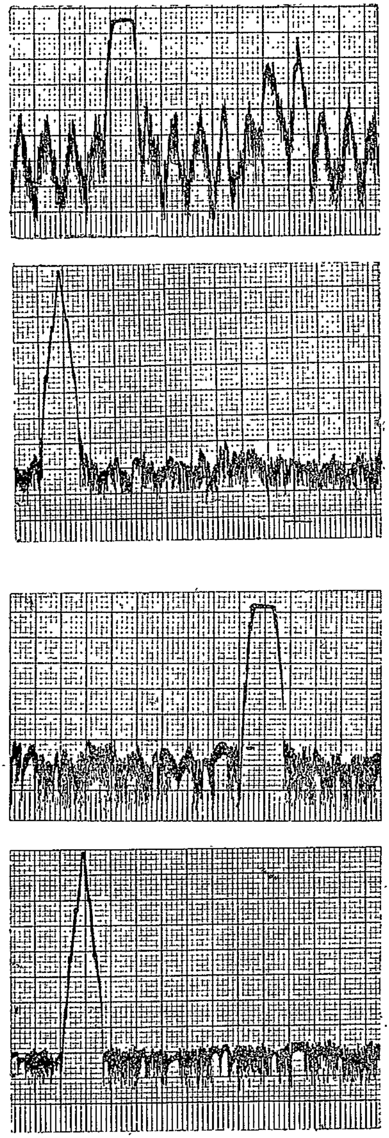
G/G_o = 1

β_o = 200 Hz

G/G_o = 8

β_o = 200 Hz

G/G_o = 1



Effect of Carrier Bandwidth on Code Correlation

FIGURE 15

the expression

$$\beta = \frac{\beta_o}{3} \left(1 + 2 \frac{G}{G_o}\right)$$

where

β = actual noise bandwidth

β_o = design noise bandwidth

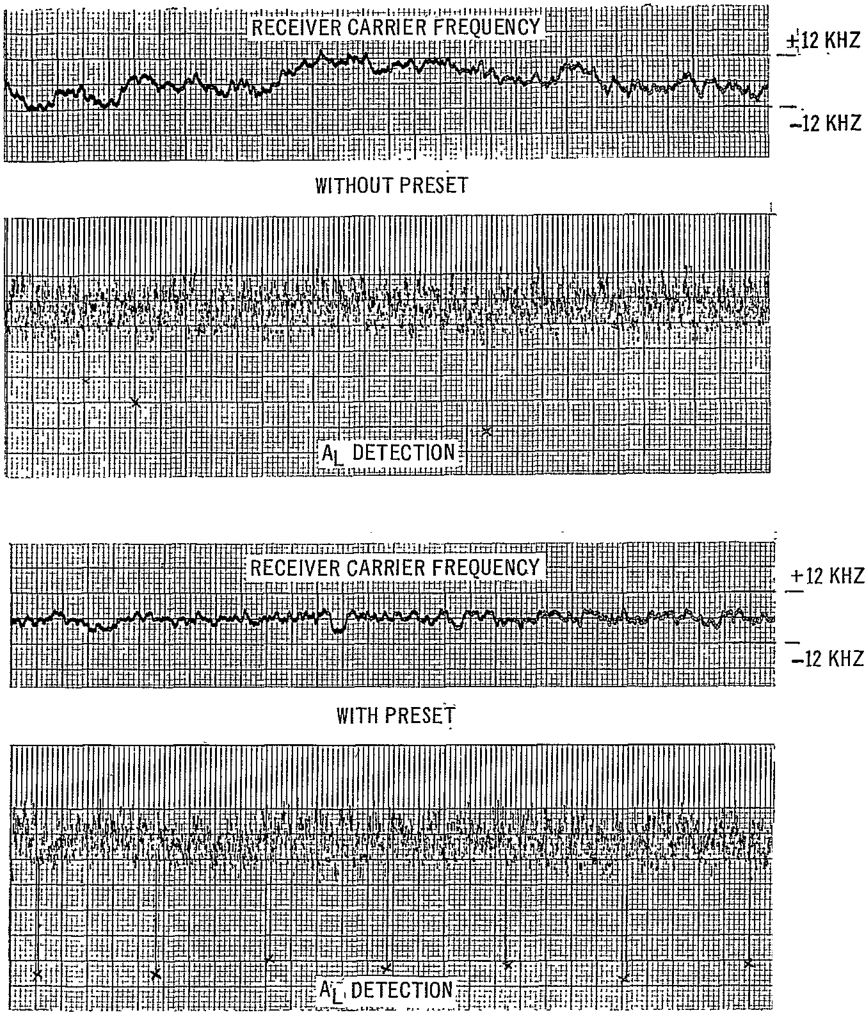
G = actual loop phase gain

G_o = design loop phase gain

These recordings were measured by opening the receiver range loop and permitting the receiver and transmitter codes to drift through synchronization. The high peaks of correlation in each case correspond to code synchronization, and the remainder of the graph is unwanted noise.

3.2.2 Preset and Gain Control Effects

The effect of the frequency preset, which during L-code acquisition holds the mean carrier loop to center frequency, is more to enhance the probability of acquisition than to decrease the code noise. The preset voltage does, however, reduce the loop phase gain during initial acquisition and prevent large frequency fluctuations. The recordings of Figure 16 show the code noise and the frequency of code capture with and without the frequency preset loop connected. For these tests, the V-2 acquisition signal was disabled so that the code search was not terminated. The code was then



x indicates detection of synchronization

PRE-SET EFFECTS UPON L-CODE SEARCH

FIGURE 16

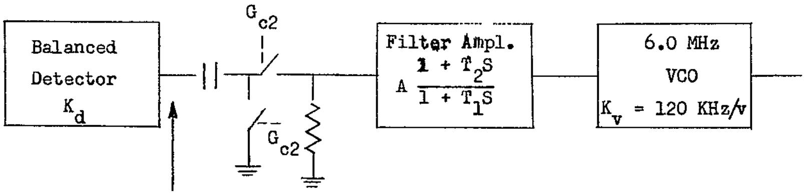
continuously shifting at 50 bits/sec, and synchronization should be detected once each code word or 2.54 seconds. It can be seen that without the preset loop, the carrier loop frequency is driven by modulation tracking over about ±12 KHz. If the frequency is far from center when code synchronization occurs, the loop capture time exceeds 20 ms and detection of synchronization is not successful. The application of the preset loop constrains the carrier frequency, ensuring that the capture time will be small.

The final precaution against modulation tracking by the carrier loop under strong signal acquisition is the application of non-coherent AGC to the receiver to limit the expansion of the carrier loop bandwidth. This AGC is developed by rectification of the apparent code noise during code search and is applied through the normal AGC system. The effectiveness of this in reducing the code noise is shown in Figure 17 under strong signal search conditions.

3.2.3 Design Summary

The final design of the carrier loop is summarized on Figure 18. The gating sequence of G_{C1} and G_{C2} is in accordance with the basic receiver timing diagram shown on Figure 7, where G_{C1} is the positive going portion of G_C and G_{C2} is the negative going portion.

The configuration of these gates is that of a basic



AGC Clamp Level = 0.5 v peak

Threshold Level = 0.25 v peak

V-1 & V-2	V-3	V-4
Optimization Level - 127 dbm $K_d = \frac{0.5}{57.3} = 8.6$ mv/deg	Optimization Level - 130 dbm $K_d = \frac{0.25}{57.3} = 4.3$ mv/deg	Optimization Level - 130 dbm $K_d = \frac{0.25}{57.3} = 4.3$ mv/deg
Gate Gain = Kg = 3/2 Duty Cycle = $\tau/T_1 = 1/2$ A = 64	Gate Gain = Kg = 3/2 Duty Cycle = $\tau/T_1 = 1/4$ A = 64	Gate Gain = Kg = 3/2 Duty Cycle = $\tau/T_1 = 1/8$ A = 64
Open Loop Transfer $H_o = K_d \times 360 \times Kg \frac{\tau}{T_1} AH_f \frac{K_v}{S}$ $= \frac{1.8 \times 10^7}{S}\left[\frac{1 + .00125S}{1 + 15.75S}\right]$	Open Loop Transfer $H_o = K_d \times 360 \times Kg \frac{\tau}{T_1} AH_f \frac{K_v}{S}$ $= \frac{4.5 \times 10^6}{S}\left[\frac{1 + .00375S}{1 + 15.75S}\right]$	Open Loop Transfer $H_o = K_d \times 360 \times Kg \frac{\tau}{T_1} AH_f \frac{K_v}{S}$ $= \frac{2.25 \times 10^6}{S}\left[\frac{1 + .00375S}{1 + 15.75S}\right]$
Closed Loop Transfer $H = \frac{H_o}{1 + H_o}$ $= \frac{1 + 1.25 \times 10^{-3}S}{1 + 1.25 \times 10^{-3}S + 8.75 \times 10^{-7}S^2}$	Closed Loop Transfer $H = \frac{H_o}{1 + H_o}$ $= \frac{1 + 3.75 \times 10^{-3}S}{1 + 3.75 \times 10^{-3}S + 3.5 \times 10^{-6}S^2}$	Closed Loop Transfer $H = \frac{H_o}{1 + H_o}$ $= \frac{1 + 3.75 \times 10^{-3}S}{1 + 3.75 \times 10^{-3}S + 7 \times 10^{-6}S^2}$
$\omega_n = 1.07 \times 10^3$ $\xi = 0.7$	$\omega_n = 535$ rad/sec $\xi = 1.0$	$\omega_n = 379$ rad/sec $\xi = 0.7$
Noise Bandwidth $\beta_L = \frac{1}{2\pi}\int_0^\infty HH^* d\omega = 565$ Hz	Noise Bandwidth $\beta_L = \frac{1}{2\pi}\int_0^\infty HH^* d\omega = 333$ Hz	Noise Bandwidth $\beta_L = \frac{1}{2\pi}\int_0^\infty HH^* d\omega = 200$ Hz

CARRIER LOOP DESIGN

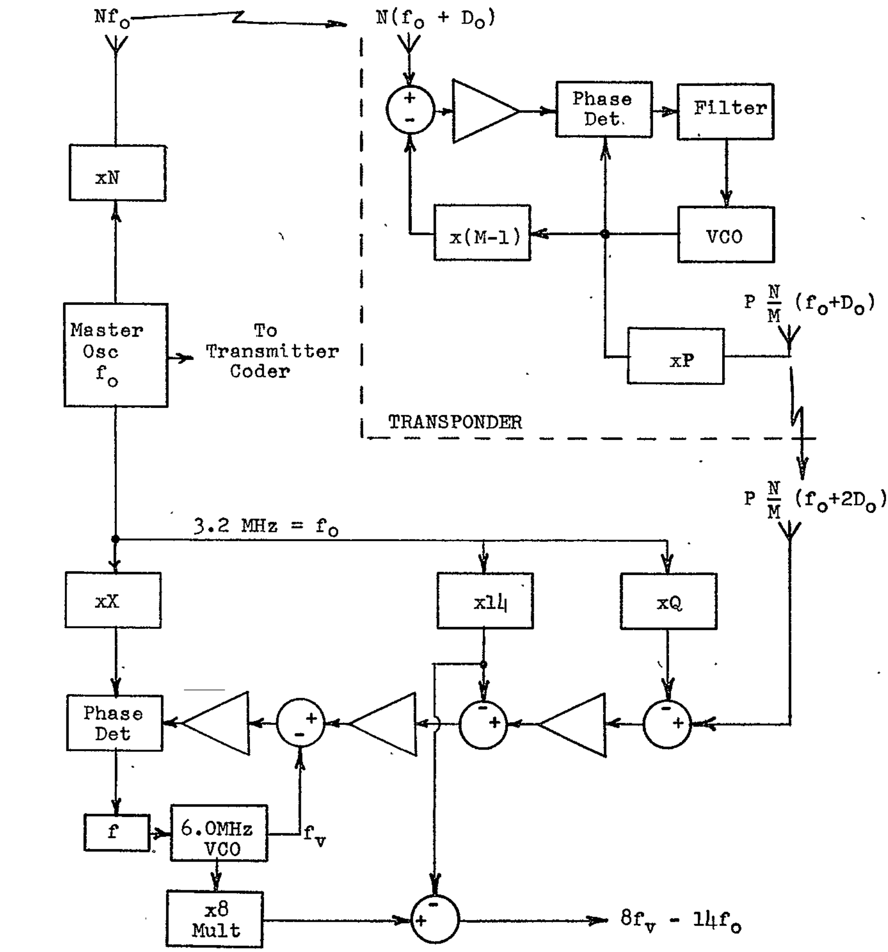
FIGURE 18

voltage doubler, although the necessity of providing a relatively constant output impedance required heavy loading of the circuit. The gain of this final detector is thus dropped from 2 to 3/2 by resistive loading. This gain is further decreased by the duty cycle of gate G_{C2} resulting in a predictable change in carrier loop phase gain as the receiver is programmed through the acquisition states.

This change in loop gain is used to perform the bandwidth switching between acquisition states. As the loop natural frequency ω_n is related to the filter lag and the loop phase gain by $\frac{1}{\omega_n} = \sqrt{\frac{T_1}{G}}$ it is necessary only to modify the filter lead between states V-2 and V-3. This is particularly desirable in preventing switching transients from perturbing the loops.

3.2.4 Doppler Extraction and Frequency Preset

One of the functions of the receiver is to extract the two-way Doppler shift of the S-band carrier frequency for measurement of the relative velocity between the Vehicle and the Transponder Station. The mechanization of the receiver is such that this entire frequency shift is contained in the frequency of the 6.0 MHz VCO in the carrier tracking loop. The VCO output is then multiplied by eight to increase the resolution of the measurement and is then heterodyned down to 3.2 MHz for delivery to the velocity extraction circuitry.



$$f_v = P\,\frac{N}{M}\,(f_o + 2D_o) - (Q + x + 14)\,f_o$$

$$8f_v - 14f_o = f_o + 8P\,\frac{N}{M}\,(2D_o) = f_o + 8(2D_P\,\frac{N}{M}\,f_o)$$

where $8(P\,\frac{N}{M} - Q - x - 14) - 14 = 1$ for phase lock

Doppler Extraction

Figure 19

This method of recovering the two-way Doppler shift of the carrier is shown in simplified form in Figure 19.

As the output signal to the velocity extraction unit is an accurate measure of the center frequency to which the receiver is tuned, comparison of this frequency against the master oscillator frequency generates the error signal for frequency preset. At the time when initial acquisition is initiated, the signal at the receiver input should contain no Doppler shift, and the output frequency of the Doppler mixer should be precisely f_o.

The measurement of the receiver error is accomplished by digital comparison of these two signals resulting in a positive voltage if the receiver frequency is low and a negative voltage if the frequency is high.

A detailed analysis of this mechanization, and a simulation of the receiver capture characteristics, is contained in Technical Memorandum M-33 dated April 15, 1965.

3.3 MODULATION LOOP DESIGN

The modulation loop derives the receiver reference code and maintains synchronization with the received code. It also performs near optimum filtering of the range information prior to presenting it to the Range Extraction Unit. The loop is strongly Rate-Aided from the carrier loop to permit the use of very narrow bandwidths.

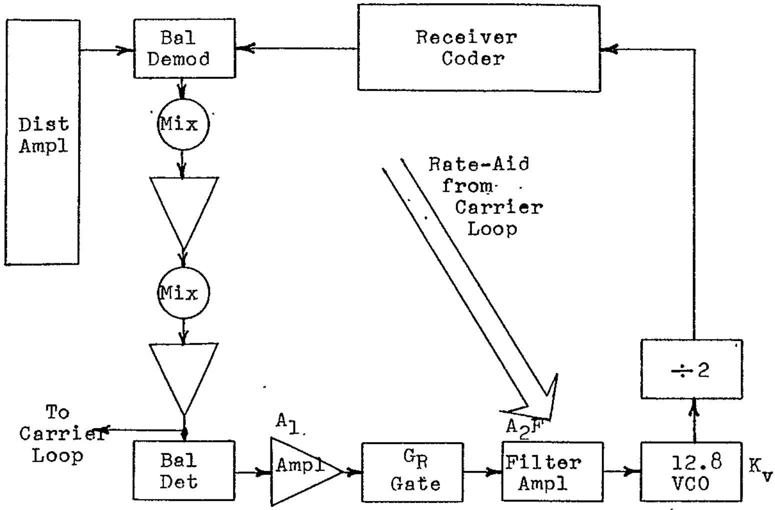
3.3.1 Bandwidth Selection

The bandwidth requirements for the modulation loop are much more easily derived than those of the carrier loop. The received signal dynamics on the modulation are low, as they are simply the Doppler shifts and rates which are associated with the modulation frequencies.

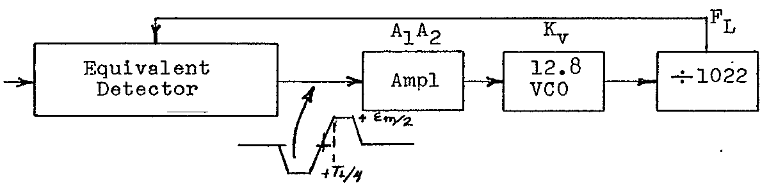
The modulation loop has two operating modes. The first of these is the acquisition mode in which the L-code is tracked. The loop remains in this mode in acquisition States V-1, V-2, and V-3. The other mode is the final Track mode which occurs in state V-4. The equivalent block diagrams of the loop in these two operating modes is shown in Figure 20.

The derivation of the error sensing curves was outlined in Section 2 of this report and was shown on Figures 4, 5, and 6. These curves pertain to the output of the equivalent detectors which include the effect of the G_R sampling gate.

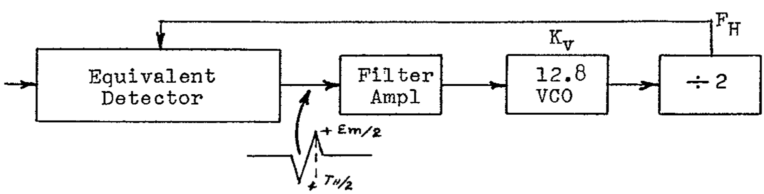
The range loop is very strongly Rate-Aided from the carrier tracking loop to almost completely remove the requirements of dynamic tracking from the loop design. This Rate-Aid mechanization will be described in Section 3.4 and its effect is to significantly reduce the tracking requirements. The signal dynamics in the modulation loop, as seen by the



Basic Modulation Tracking Loop



Fquivalent Loop Tracking L-Code



Equivalent Loop Tracking H-Code

Modulation Tracking Loop

FIGURE 20

equivalent detector, are tabulated below.

Acquisition State	Maximum Doppler Shift		Maximum Doppler Rate	
	Without Rate-Aid	With Rate-Aid	Without Rate-Aid	With Rate-Aid
V-1	±1.0 Hz	±1.0 Hz	.042 Hz/sec	.042 Hz/sec
V-2	±1.0 Hz	±1.0 ⟶ ±.0029 Hz	.042 Hz/sec	.042 ⟶ .0001
V-3	±1.0 Hz	±.0029 Hz	.042 Hz/sec	.0001 Hz/sec
V-4	±511 Hz	±1.0 Hz	21.3 Hz/sec	.006 Hz/sec

It is seen that although the Rate-Aid is ineffective while operating with the Reversed Doppler in the Transponder Station, it becomes effective in state V-2. The signal dynamics which remain after Rate-Aid are small and do not constrain the selection of the loop responses.

The selection of the loop bandwidth is almost entirely dictated by the filtering, or data smoothing requirement. As the final data output occurs at 4 readings/second, it is desirable that each sample be smoothed to provide a near optimum estimation of range delay. This smoothing corresponds to a loop low pass noise bandwidth of the order of 4 Hz. Comparison of this bandwidth with the residual Doppler shift shows that the Range Loop may be first order in states V-1, V-2, and V-3, but must be second order in state V-4.

The application of the strong Rate-Aid also relieved the range loop of the high gain requirements to track the Doppler extremes. The loop lag will be the greatest in V-1 and will be about 24 degrees for a first order loop with 4 Hz noise bandwidth. The lag in tracking state V-4, will not be as great, but the lag will contribute directly to a range measurement error.

3.3.2 Drift Effects

Although the signal to be tracked has been greatly reduced by the Rate-Aid, oscillator drift and amplifier bias effects must be tracked out with low lag. Assigning a net bias of $\pm\epsilon d$ to the output of the equivalent detector, where ϵd includes the net effect of all DC drifts, and assuming that the 12.8 MHz oscillator drift is $\pm\Delta f$, the range error will be

$$\Delta = \left(\pm\epsilon d \pm \frac{\Delta f}{A_1 A_2 K_v}\right) \frac{T_H/2}{\epsilon m/2} \text{ seconds}$$

$$= \left(\pm\epsilon d \pm \frac{\Delta f}{A_1 A_2 K_v}\right) {T_H}/{\epsilon m} \text{ seconds}$$

The range loop gain at threshold in state V-4 will be

$$G_4 = \frac{\epsilon m/2}{T_H/2} \left(\frac{v}{sec}\right) \times A_1 A_2 \left(\frac{v}{v}\right) \times \frac{K_v}{2} \, \frac{\text{cycles}}{\text{v - sec}} \times T_H \, \frac{\text{sec}}{\text{cycle}}$$

$$= \frac{\epsilon m A_1 A_2 K_v}{2} \text{ sec}^{-1}$$

The ranging error due to drift effects in the modulation loop will be

$$\Delta = \left(\pm \frac{\epsilon d}{\epsilon m} \pm \frac{\Delta f}{2G_4}\right) T_H \text{ seconds}$$

and as $T_H = 0.156\ \mu sec$

$$\Delta = 23.4 \left[\pm \frac{\epsilon d}{\epsilon m} \pm \frac{\Delta f}{2G_4}\right] \text{ meters range error}$$

It can be seen that the loop gain is ineffective in reducing range error due to amplifier or detector drifts, but does directly reduce error from oscillator center frequency stability. Allocating a peak error from each of these sources of 0.2 meters at the receiver threshold,

$$0.2 > 23.4 \frac{\Delta f}{2G_4} \qquad\qquad 0.2 > 23.4 \frac{\epsilon d}{\epsilon m}$$

Let $\Delta f = \pm 64$ Hz (5 ppm) Let $\epsilon m = 0.25$v (threshold)

$$\therefore G_4 > 3600 \qquad\qquad \epsilon d < \pm 2.1 \text{ mv}$$

At the nominal AGC clamp level, the detector gain will be increased by a factor of two, or $\epsilon m = 0.5$v, further reducing these drift effects under normal operating conditions.

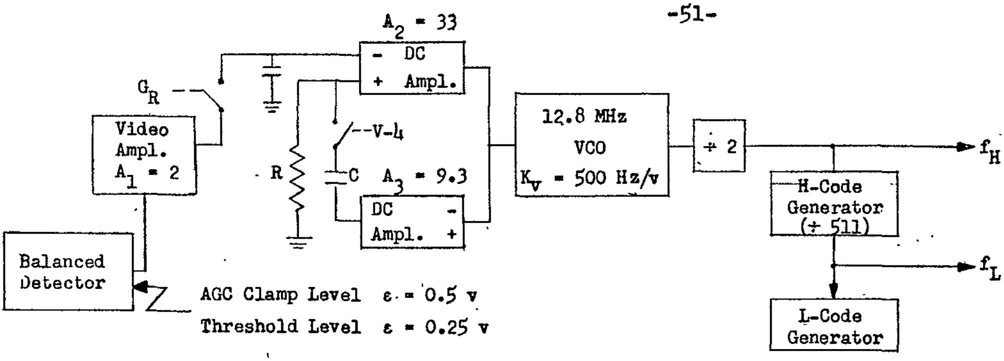
The selected gating method derived in Section 2 of this report shows that there is an automatic increase in detector gain by a factor of $\frac{f_H}{2f_L}$ or 256:1 when tracking is switched from L-code to H-code in state V-4. This permits the use of a type one first order loop in V-1, V-2, and V-3 with a noise bandwidth of 4 Hz. This requires that

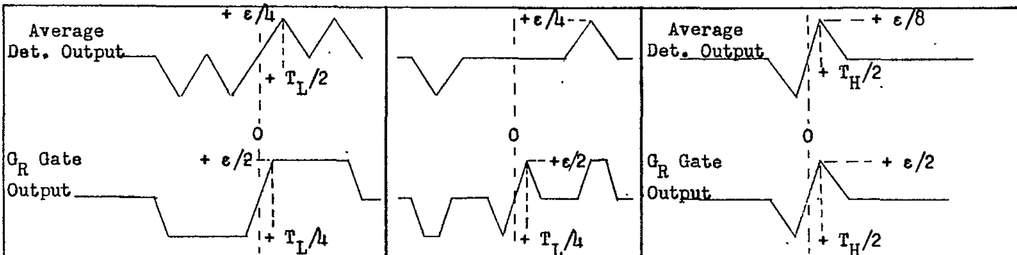
$$G_{1,2,3} = 16 \text{ sec}^{-1}$$

The gain will then increase to $G_4 = 16 \times 256 = 4100 \text{ sec}^{-1}$ in V-4 satisfying the gain constraint. The detail design of this modulation loop is summarized in Figure 21, and the measured performance is shown on Figures 22 and 23.

3.4 RATE-AID LOOP DESIGN

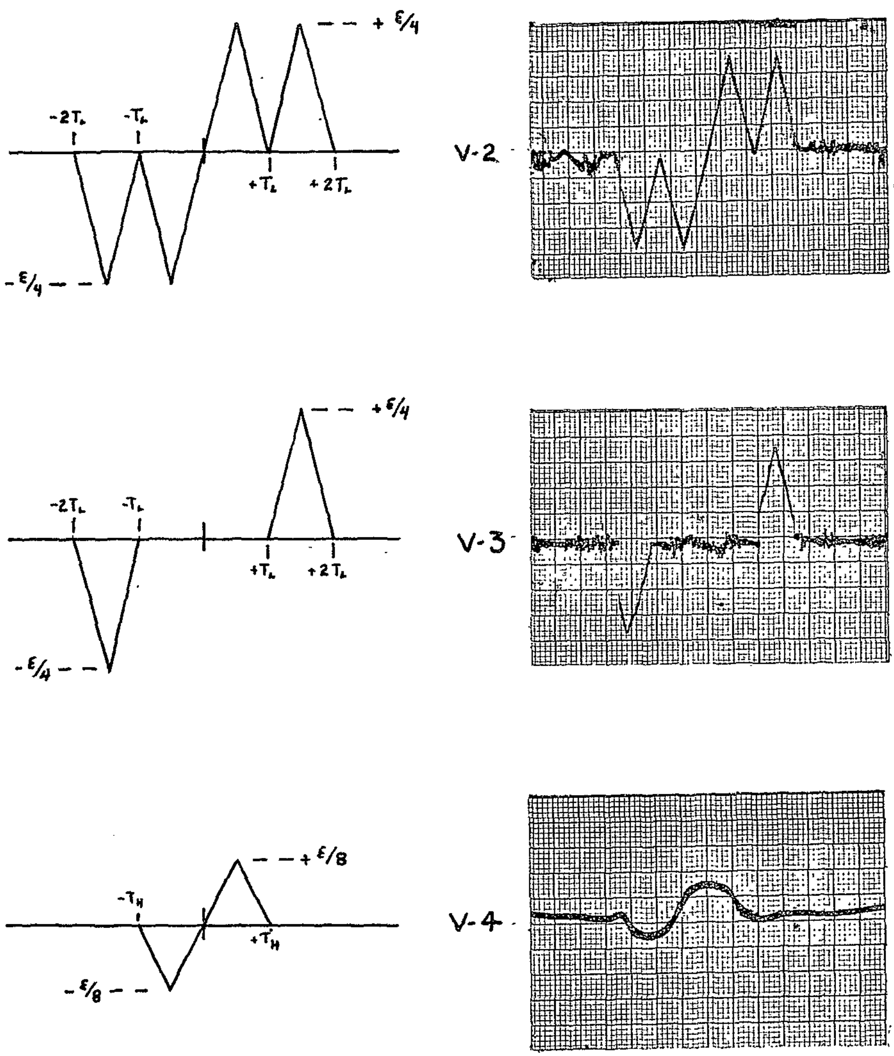
The receiver velocity tracking (carrier) loop and the range tracking (modulation) loop normally operate independently. There is a small degree of cross-coupling due to tracking lags in the loops as was shown in Section 2, but these cross-coupling coefficients have deliberately been made small. It was shown in Section 3.3 that the Doppler shifts and the Doppler rates which the modulation loop must track are those shifts and rates associated with the modulation frequencies. The receiver is intended to operate with relative velocities up to ±12,000 m/sec and accelerations up to ±400 m/sec^2. These dynamics are not compatable with the desired smoothing function of the modulation loop corresponding to a low pass bandwidth on the order of 4 Hz. The modulation tracking loop gain would be forced to be in excess of 6×10^4, requiring a filter time constant on the order of 1000 sec, and the acceleration lags would exceed 100 degrees with corresponding loss of tracking capability.





Open Loop Transfer $H_o = \frac{\varepsilon/2}{T_L/4} \times 2 \times 33 \times \frac{500}{2S} \times \frac{1}{511}$ $= 64 \frac{\varepsilon}{S}$ $\varepsilon = 1/4$ v $H_o = \frac{16}{S}$	Same as V-1 ε V-2	Open Loop Transfer $H_o = \frac{\varepsilon/2}{T_H/2} \times 2 \times 33 \times \frac{1 + RCS}{1+(A_2A_3+1)RCS} \times \frac{500}{2S}$ $H_o = 16500 \frac{\varepsilon}{S} \left[\frac{1 + RCS}{1 + 307\ RCS}\right]$ $\varepsilon = 1/4$ v RC = 0.15 sec. $H_o = \frac{4125}{S} \left[\frac{1 + 0.15S}{1 + 46S}\right]$
Closed Loop Transfer $H = \frac{H_o}{1 + H_o}$ $= \frac{1}{1 + S/16}$	Same as V-1 ε V-2	Closed Loop Transfer $H = \frac{H_o}{1 + H_o}$ $= \frac{1 + 0.15S}{1 + 0.15S + .0111S^2}$
Noise Bandwidth $\beta_L = \frac{1}{2\pi} \int_0^\infty HH^* \, d\omega = 4\text{Hz}$	Same as V-1 ε V-2	Noise Bandwidth $\beta_L = \frac{1}{2\pi} \int_0^\infty HH^* \, d\omega = 5\text{Hz}$ $\xi = 0.7$

RANGE LOOP DESIGN
FIGURE 21

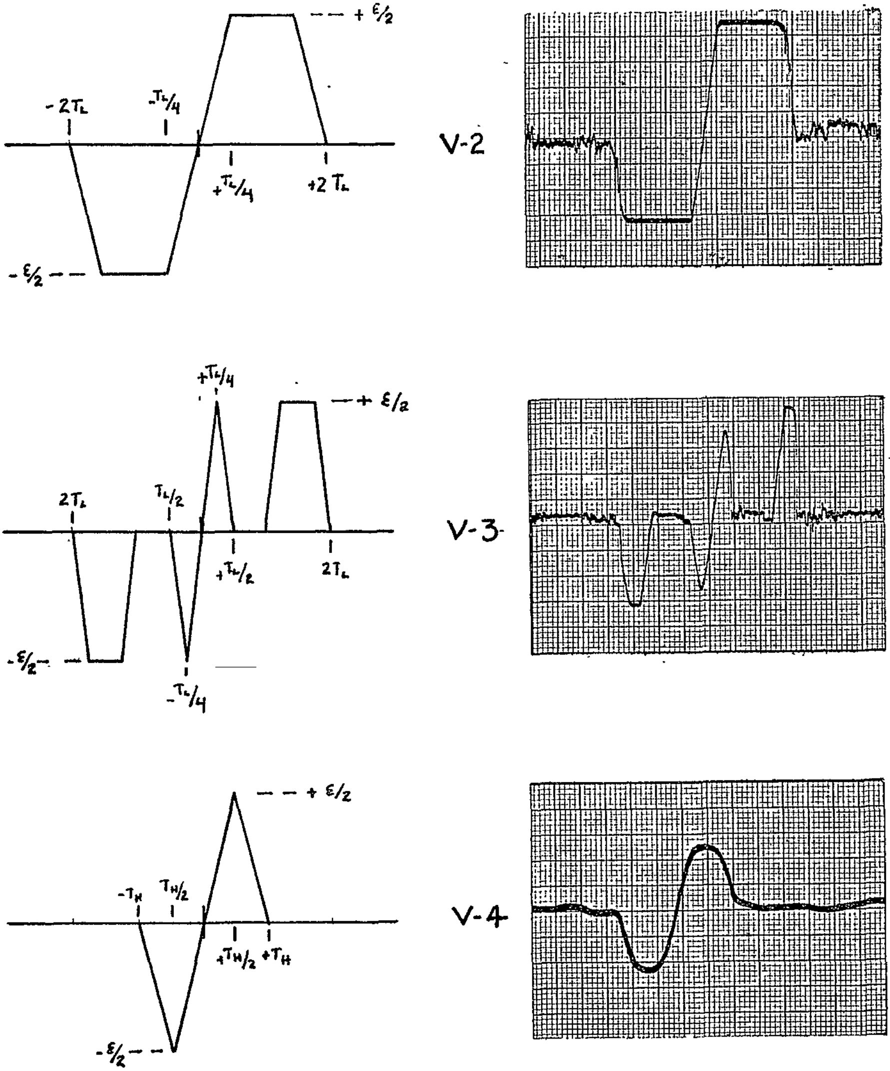


Calculated

Measured

Range Code Correlation
Detector Output

FIGURE 22



Calculated Measured

Range Code Correlation
Range Gate Output

FIGURE 23

The carrier tracking loop must also track the signal dynamics, but as it has no corresponding smoothing function in the loop design it has a relatively wide bandwidth and is able to track properly. This permits the use of carrier to modulation Rate-Aid in which the Doppler effects on the carrier are scaled down and used to program the modulation tracking loop. An extremely effective Rate-Aid is possible in the AROD receiver design due to the basic coherence between the received carrier frequency and the frequency of the modulation on that carrier. Referring to Section 3.2.4, Figure 19, it is seen that the carrier and the modulation are both derived from the same frequency source and are synthesized coherently. The ratio between the frequency of the 6.0 MHz voltage controlled oscillator and the 12.8 MHz voltage controlled oscillator is

$$\frac{f_v \text{ carrier}}{f_v \text{ mod}} = \frac{\left(P\,\frac{N}{M} - Q - X - 14\right) f_o + P\,\frac{N}{M}\,(2\,D_o)}{4f_o + 4(2\,D_o)} .$$

Dividing the 6.0 MHz oscillator output by the factor $\frac{PN}{4M}$ will yield the synthesized frequency

$$4\left|1 - \frac{M}{PN}\,(Q + X + 14)\right| f_o + 4(2\,D_o)$$

which contains precisely the expected Doppler shift of the 12.8 MHz oscillator in the modulation tracking loop superimposed on a frequency offset f_{RA}, where

$$f_{RA} = 4\left[1 - \frac{M}{PN}\,(Q + X + 14)\right] f_o$$

To satisfy the coherence of the Doppler extraction system when there is no Doppler shift, from Figure 19

$$8f_{v\ carrier} - 14f_o = f_o$$

$$8f_{v\ carrier} = 15f_o = 8(P\frac{N}{M} - Q - X - 14)\ f_o$$

$$\left[1 - \frac{M}{PN}(Q + X + 14)\right] f_o = \frac{15}{8}\frac{M}{PN} f_o$$

The offset frequency for the Rate-Aid signal is then

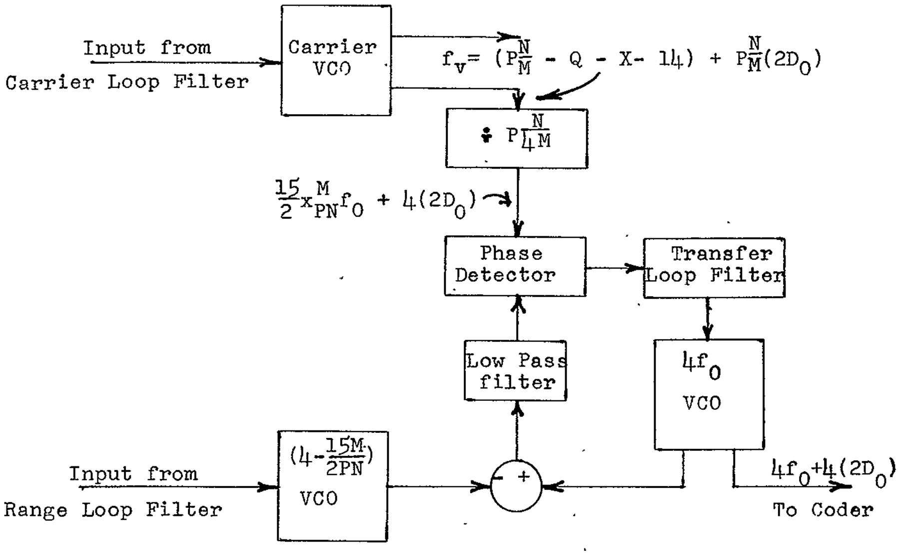
$$f_{RA} \doteq \frac{15}{2}\frac{M}{PN} f_o$$

This signal is applied to the modulation tracking loop as shown in Figure 24.

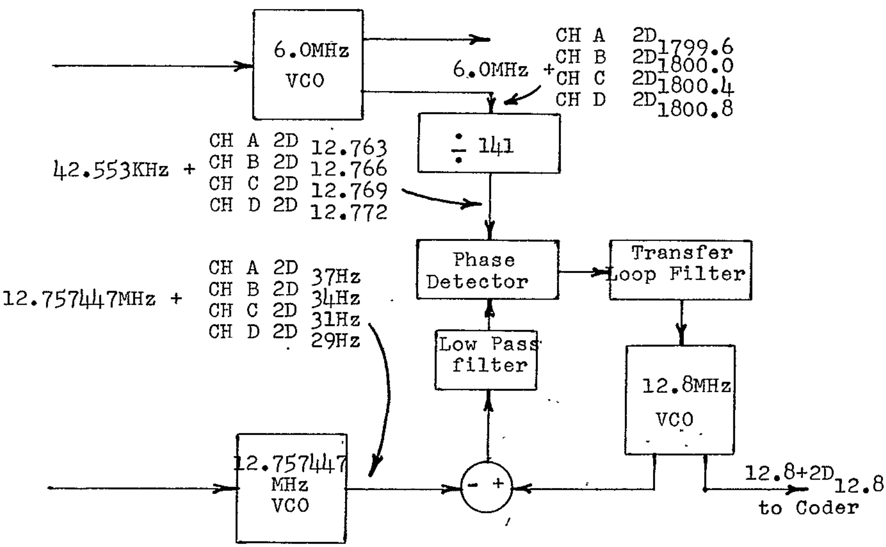
This Rate-Aid transfer loop replaces the 12.8 MHz oscillator shown in the modulation loop design. The transfer loop bandwidth has been made wide with respect to the modulation loop response so the effects upon the tracking can be expressed in terms of the phase of the carrier oscillator (θ_c), the phase of the modulation loop oscillator (θ_v), the resultant phase of the range oscillator (θ_R), and the transfer loop response Y(s).

$$\theta_R = (\theta_v + \frac{\theta_c}{141})\ Y(s)$$

It can be seen that the noise on the ranging signal phase is increased insignificantly by the Rate-Aid due to the



A IDEAL RATE - AID



B APPROXIMATE RATE - AID

FIGURE 24 RATE AID TRANSFER LOOP

high division ratio. The response of the transfer loop is that of a 500 Hz low pass noise bandwidth where

$$Y(s) = \frac{1 + .0016S}{1 + .0016S + 1.12 \times 10^{-6}S^2}$$

which is a second order loop with

$$\omega_n = 945 \text{ rad/sec}$$

$$\xi = 0.7$$

$$\beta_L = 500 \text{ Hz}$$

In the range of frequencies up to about 100 Hz, the response of this transfer loop to an error signal in the modulation loop is very nearly unity. Thus, the entire transfer Rate-Aid loop appears to be an equivalent 12.8 MHz VCO with the gain constant (and frequency stability) of the 12.757447 MHz VCO.

4.0 RECEIVER INTEGRATION

The final block diagram of the Vehicle Tracking Receiver, showing only one of the four receiving channels, is shown in Figure 25. This diagram is oriented toward the physical construction of the receiver as each block is one of the modular assemblies of the receiver. The terms "Carrier Tracking Loop," "Modulation Tracking Loop," and "Receiver Coder," as used on this diagram, refer to the major subassemblies of the receiver and include assorted bits and pieces in addition to the loops as discussed in Section 3 of this report. The detailed circuit design, the methods of implementing the switching functions, and discussions of circuit performance will be reserved for an additional report.

The receiver as a complete unit has several constraints upon the design which result from the effects of the multiple receiving channels, and are not directly related to single channel performance. These include the effects of intermodulation, adjacent channel jamming, and the spectral purity requirements of the reference signals supplied from the frequency synthesizer.

4.1 INTERMODULATION AND JAMMING

The four receiving channels are not widely separated in frequency. There is a nominal 400 KHz channel spacing, but

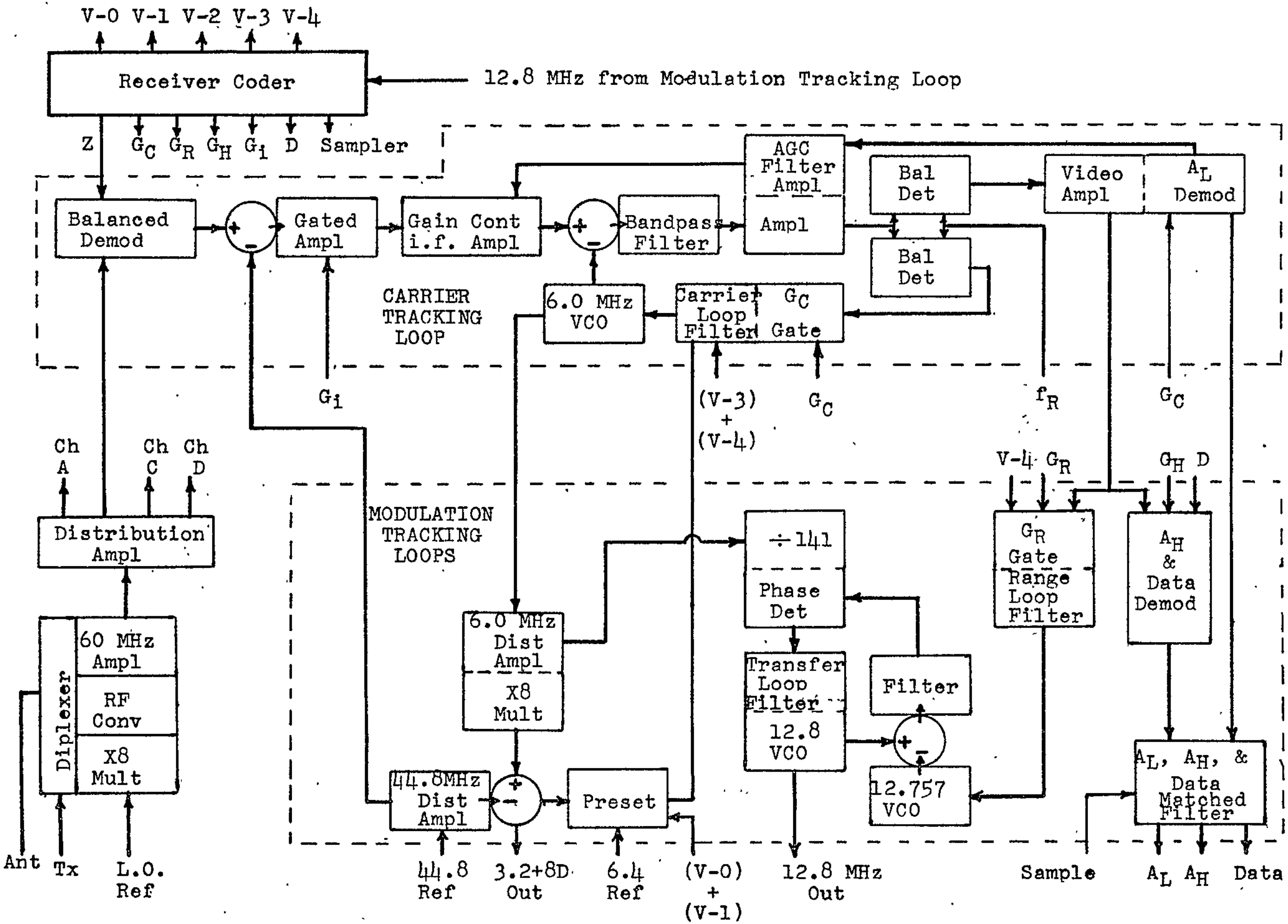


FIGURE 25 Vehicle Tracking Receiver

this is modified by the range of Doppler shifts which cause each of the incoming signals to be independently variable over a ±140 KHz range. As the receiver signals are biphase modulated by the transmitter coders, the resultant spectra will occupy a wide bandwidth with an envelope of the form $\frac{\sin^2 x}{x^2}$. The signals from each responding station will then overlap each other and power from each transponder will lie in-band to every receiving channel.

There are two major effects from this adjacent channel interference. The first of these is the production of intermodulation products and the second is outright jamming of a channel by another due to power level differences.

4.1.1 Intermodulation Effects

Direct intermodulation of the received codes, or multiple products of one code with another code can degrade the receiver performance. As the codes are all of the same sequence, and differ only in their phases, products will be of the form of the original code at a new phase modulated upon the difference frequency. The difference frequency will be 400 KHz, 800 KHz or 1200 KHz as the channel spacing is a uniform 400 KHz and there are four channels. This can in turn reflect into a third channel with a modulation which can be correlated and cannot be discarded by subsequent narrow band filtering.

Saturation effects will have a similar behavior. The second harmonic of a signal will be formed which is not suppressed carrier modulated. This will mix with another channel signal reflecting into a third channel. The signal levels and power handling capabilities of all stages prior to demodulation and channel separation have been carefully arranged to provide 70 db protection against these potential interference sources.

4.1.2 Adjacent Channel Jamming

Limiting occurs in each receiver channel with an output level of 2.0 volts peak. The loops become underdamped with the levels less than 0.25 volt peak in V-3 or V-4 and 0.5 volt peak in V-1 or V-2. The linear region is then either 18 db or 12 db and the receivers can tolerate these jamming to signal power ratios in the i.f. channel. As intermodulation and other inadvertent products between signals have been controlled, the primary effect of adjacent channel jamming will be suppression in the limiter of the desired signal by the jammer. The degree of protection against this effect depends upon the amount of adjacent channel power which can enter one of the tracking receiver channels.

Both the desired signal and the adjacent channel interference will be of the same form, that is the transmissions will be biphase-modulated pseudonoise coded signals.

The codes may be either the Low code alone or the Low and High codes together. The code sequences on each of the transmissions are identical although the phases are dependent upon the particular slant ranges and are arbitrary.

There are a number of combinations of the form of the interferring signal and the particular reference code in the receiver channel which affect the amount of interference power which can enter the receiver. These combinations are:

(1) Interferring signal is L-code
Receiver reference is $L \oplus f_L$
Interferring and tracking codes are synchronous

(2) Interferring signal is L-code
Receiver reference is $L \oplus f_L$
Interferring and tracking codes are non-synchronous

(3) Interferring signal is L-code and H-code
Receiver reference is $L \oplus f_L$

(4) Interferring signal is L-code
Receiver reference is $L \oplus f_L$ and $L \oplus f_H$

(5) Interferring signal is L-code and H-code
Receiver reference is $L \oplus f_L$ and $L \oplus f_H$

In all cases the interferring signal is assumed to be the adjacent channel and separated by 400 KHz from the signal which is being tracked. The bandwidth through which the interferring signal can enter the tracking channel is that of a single pole filter with a 3 db bandwidth of 125 KHz corresponding to a noise bandwidth of 200 KHz. Although the transmitted signals appear to be broad band noise, they are actually composed of a large number of spectral lines separated by about 50 Hz. The interference power which will affect the tracking channel will then be of the form of a number of these spectral lines.

In the particular case where the transmitter signal is L-code, the receiver is tracking L-code, and the codes are synchronous because the ranges to the desired station and interferring station are identical, the receiver reference code will demodulate the interference resulting in a 12.5 KHz square wave modulated signal 400 KHz off center frequency. The power in each component will be $\frac{8J}{n^2\pi^2}$ (n odd) where J is the interference power and n is the order of the side-band. The noise bandwidth of the predetection filter will be about 200 KHz so sidebands from n = 25 to n = 39 will be effective. This power will be about 22 db below the total interference power. As the receiver is tracking L-code, it can tolerate a 12 db interference to signal ratio resulting in a 34 db protection against this combination.

When both of the transmissions are L-code, but the codes are not in synchronism, the resultant interference spectrum out of the balanced demodulator will include the cross product between the interference code and the reference code with spectral components spaced at 50 Hz intervals. The total power in the 300 KHz to 500 KHz region will be nearly the same as if the codes were correlated even through more components are involved resulting in about 34 db protection against non-synchronous L-code interference.

When either but not both of the transmissions are L-code and H-code, one half of the power will be in the H-code and one half in the L-code. The effective interferring power will be $\frac{1}{2} \times \frac{200 \times 10^3}{6.4 \times 10^6}$ or 18 db below the total power from the H-code and 25 db below the total power from the L-code. The effective suppressing power is then 17 db below the interference level giving 29 db protection against H-code interference while tracking L-code.

When the receiver is tracking L-code and H-code and the interference is also L and H, the result will be 3 db higher than if one of the signals were L-code alone for all of the power is spread uniformly through the region of interest. The receiver, however, is able to tolerate a 6 db higher interference level when tracking in V-4. The protection against this interference mode will then be about 32 db.

The primary region of interest for receiver operation lies between 100 and 2000 km transponder to vehicle range. The relative difference in received power will then be as great as 26 db for simultaneous operation with maximum and minimum station distances. The protection which the receiver provides is in all cases greater than the maximum expected interference to signal ratios.