ESD-TR-72-272

FEASIBILITY STUDY OF BANDWIDTH REDUCTION TECHNIQUE FOR LORETS SYSTEM

ELECTRONIC COMMUNICATIONS, INC.
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E. J. Dalabakis
G. R. Shoubridge

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13. ABSTRACT

The purpose of this feasibility study and test demonstration was to conduct analysis for a bandwidth reduction technique to be incorporated in a LORAN Retransmission (LORETS) System. The design concept resulting from the analysis was then breadboarded and tested in the laboratory to prove out the feasibility of using a bandwidth reduction technique with a LORETS System. The analysis that led to the original design concept is presented in this report. The breadboard bandwidth reduction system is also presented with results of this system with ~~(1)~~ simulated LORAN signals, ~~(2)~~ with the system hard wired at the down converted spectrum, and ~~(3)~~ with the system over a hard wired RF link using LORAN C signals. The results illustrate that bandwidth reduction and expansion is indeed feasible using the techniques developed herein and reduces the LORETS transmission bandwidth from 220 kHz to 57.5 kHz with slight degradation of the total system capability.

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E.J. DALABAKIS
G.R. SHOUBRIDGE

ELECTRONIC COMMUNICATIONS, INC.
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Approved by: K.W. Kirk
K.W. Kirk, Manager
Systems Engineering

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ABSTRACT

The purpose of this feasibility study and test demonstration was to conduct analysis for a bandwidth reduction technique to be incorporated in a LORAN Retransmission (LORETS) System. The design concept resulting from the analysis was then breadboarded and tested in the laboratory to prove out the feasibility of using a bandwidth reduction technique with a LORETS System. The analysis that led to the original design concept is presented in this report. The breadboard bandwidth reduction system is also presented with results of this system with (1) simulated LORAN signals, (2) with the system hard wired at the down converted spectrum, (3) with the system over a hard wired RF link using LORAN C signals. The results illustrate that bandwidth reduction and expansion is indeed feasible using the techniques developed herein and reduces the LORETS transmission bandwidth from 220 kHz to 57.5 kHz with slight degradation of the total system capability.

SECTION 1
INTRODUCTION

The basic LORAN signal is an amplitude modulated 100 kHz carrier. The characteristics of the LORAN modulating waveform are such that the LORAN spectrum is approximately 100 kHz ±10 kHz.

This bandwidth of 20 kHz is the basic LORAN information bandwidth. This is the only information that must be retained to determine the LORAN position. In the present LORETS scheme, the LORAN signal at 100 kHz is applied directly to the retransmission link so that the retransmission link bandwidth is 220 kHz rather than the minimum required LORAN bandwidth of 20 kHz. Unfortunately, there is no way other than coherent demodulation to reduce the bandwidth for LORETS to the minimum 20 kHz and any reasonable coherent demodulation system is nearly as complex as the full LORAN receiver. It is possible, however, to translate the LORAN band (90 to 110 kHz) down to a lower region (approximately 5 - 25 kHz), for retransmission via LORETS and translate it back to its original frequency at the LORETS output. Since the LORAN demodulation process requires measuring the 100 kHz carrier phase at a specific point within the pulse envelope, it is necessary in any retransmission scheme to preserve the phase relationship between the envelope and the 100 kHz carrier. This relationship is, in general, not preserved when a signal is subjected to arbitrary frequency translation and filtering processes. It is preserved, however, if a frequency translation is followed by a retransmission to the original frequency and if the local oscillator signals used for both the translation and the retranslation are of identical phase. A derivation of this fact is presented in Appendix A.

A functional diagram of the implementation of the LORETS bandwidth reduction scheme which was developed under this contract is shown in Figure 1-1. The received LORAN signal occupying the 90 to 110 kHz band is mixed with a

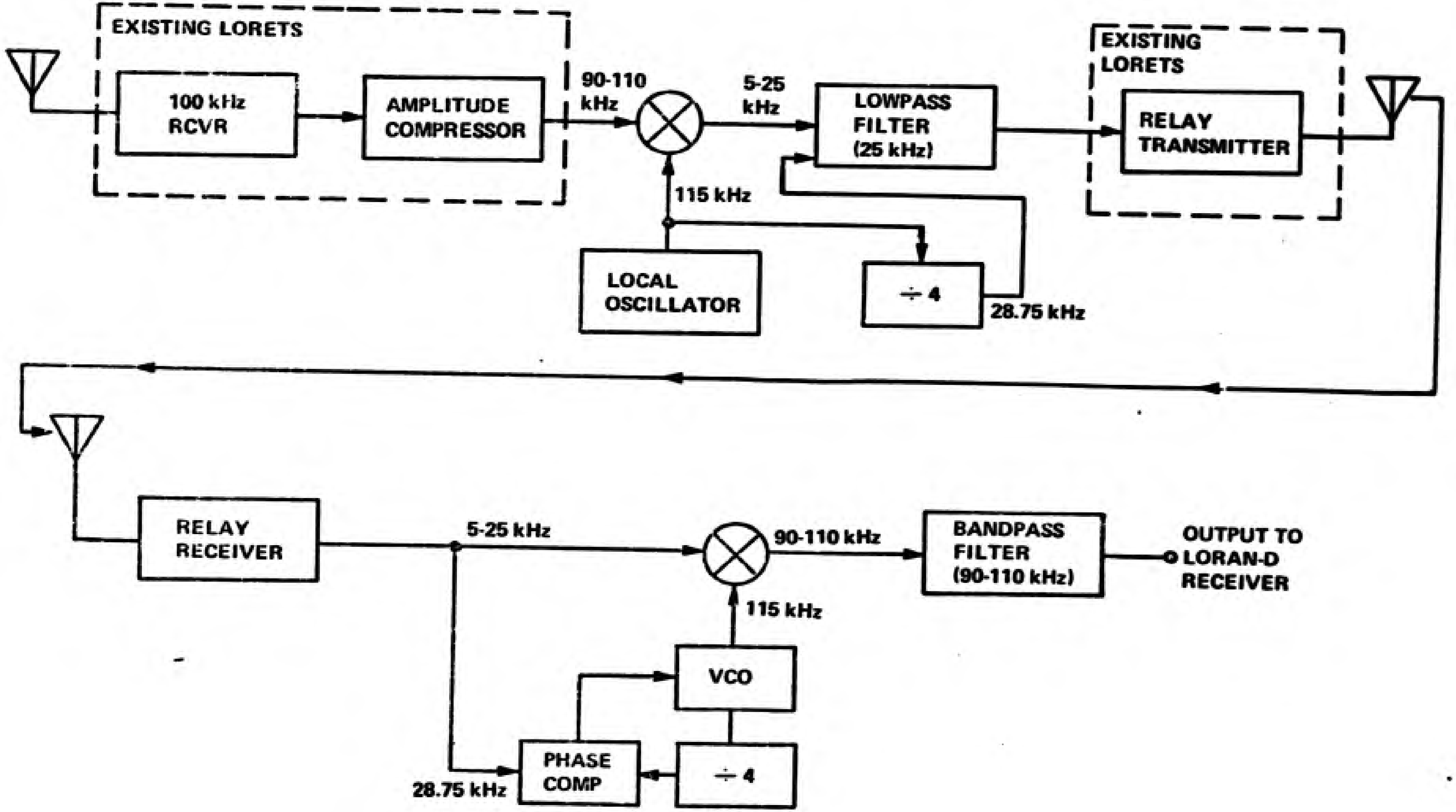


FIGURE 1-1. BLOCK DIAGRAM, LORETS BANDWIDTH REDUCTION

115 kHz local oscillator to shift it down to the 5 to 25 kHz band. A pilot signal from the local oscillator (local oscillator divided by four) is added to the down converted LORAN signal for transmission. At the receive terminal, the pilot signal is recovered by a phase lock loop and locks the 115 kHz receive local oscillator which is then used to coherently upconvert the 5 to 25 kHz signal to its original LORAN format. Following the mixer is a 90 to 110 kHz bandpass filter to select the proper sideband of the upconversion process. This signal is then in the same format with Master, Slave 1 and Slave 2 relative time relationships unchanged. This signal can then be fed to a LORAN C/D receiver to compute TDA and TDB.

SECTION 2

DESIGN CONCEPT AND THEORY OF OPERATION

The concept behind the basic approach used in the BW reduction technique is related to the fact that all the meaningful information in the LORAN pulse stream is contained in the frequency band of 90 to 110 kHz. When transmitting this information via a UHF link the minimum required bandwidth is 220 kHz. By frequency translating the 90 to 110 kHz information to range 5 to 25 kHz, a 4:1 saving in UHF bandwidth can be achieved. Past experience with FDM systems suggests that by using single sideband suppressed carrier amplitude modulation techniques, very little distortion in both amplitude and phase response is produced in the frequency translation process if a reasonable separation is selected between the carrier frequency and information band being processed. Once the information has been converted to the new bandwidth, transmitted over a UHF link and detected by conventional techniques, the information bandwidth is retransmitted to its original frequency allocation by the reverse process. To preserve frequency and phase integrity at the receiver, the carrier frequency used in the receiver retranslation process is controlled by a pilot tone derived from the carrier frequency in the transmitter which is transmitted as part of the information bandwidth.

2.1 IMPLEMENTATION OF THE BW REDUCTION TECHNIQUE

For the purpose of demonstrating the feasibility of this scheme, the following steps were followed:

a. Breadboard and test a modulator in the correct frequency range to determine the requirements of the low pass filter for the transmit path.

b. Design a low pass filter for use with this modulator and measure the in-band characteristics for amplitude and phase.

c. Repeat step a. for the receive path to arrive at the requirements for the receive filter.

d. Connect a pair of modulators and filters in tandem using the same carrier source for both modulators and check the end to end performance of the link using a signal generator and selective VTVM to measure the characteristics of the path in the frequency range of 90 to 110 kHz. From the data taken during this test and a subsequent test using the LORAN pulse train from a receiver, it was decided that the technique was feasible when a single carrier frequency source was used for both the transmit and receive paths.

Tables 2-1 and 2-3 were used for steps a. and c. to determine the filter requirements for both paths. The figures shown for attenuation in ( ) are actual measurements, those shown in [ ] are the expected values from this type of modulator. The values marked with an * represent in-band distortion products which cannot be removed by filtering and must be suppressed in the modulation process.

- TRANSMIT FILTER REQUIREMENTS

The test set-up that was used for step a. is shown in Figure 2-1 and measurements were recorded in dB below the wanted sideband with an input signal level (i.e., 90 to 110 kHz) of approximately 50 mv rms and a carrier level (i.e., 115 kHz) of 300 mv peak to peak.

A circuit diagram of the modulator and buffer stage is shown in Figure 2-2 with the results tabulated in Table 2-1.

From Table 2-1 it can be seen that interfering inband products (i.e., the asterisk values which fall in the 5 to 25 kHz band) are a minimum of 60 dB below the wanted sideband which indicates that the modulation process used is good. The closest out of band product is in the 120 to 140 kHz band and is 40 dB down and therefore would require an additional 20 dB rejection by the filter to be below the level of the inband distortion products. The upper sideband is in the 205 to 225 kHz range and would require 60 dB of rejection for the same criteria. Since this frequency is three octaves above the (5 to 25 kHz)

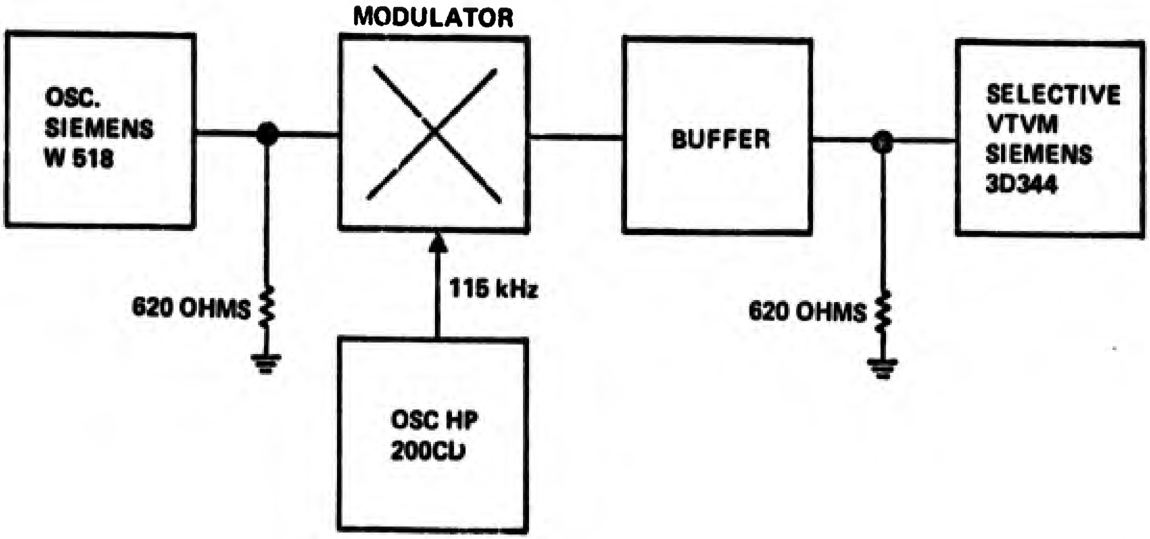


FIGURE 2-1. MODULATOR TEST SET-UP

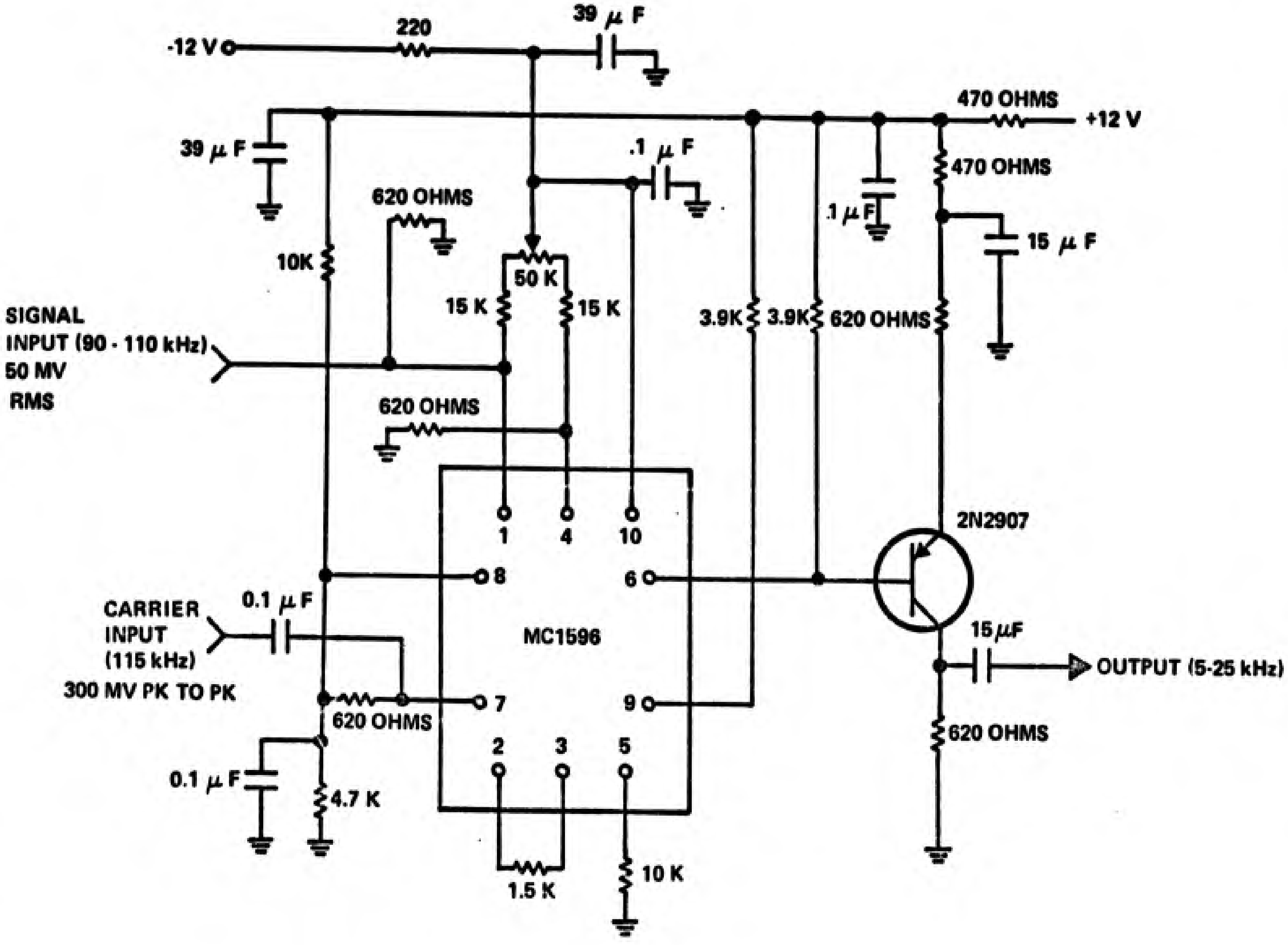


FIGURE 2-2. MODULATOR AND BUFFER CIRCUIT

TABLE 2-1.

Modulation Products (Down Conversion)

M = Modulating Frequency = 90 - 110 kHz

C = Carrier Frequency = 115 kHz

		C= 115 kHz		2C= 230 kHz		3C= 345 kHz		4C= 460 kHz	
		C± = 115 kHz		2C± 2C= 230 kHz		3C± 3C= 345 kHz		4C±4C= 460 kHz	
	f kHz	f kHz	dB	f kHz	dB	f kHz		f kHz	
M	90-110	5-25 or 205-225	[0] (0)	140-120 or 320-340	[40] (39)	225-235 or 435-455	[10] (11)	370-350 or 550-570	[40] (38)
2M	180-220	65-105 or 295-335	[60] (63)	50-10* or 410-450	[60] (62)	165-125 or 525-565	[60] (58)	280-240 or 640-680	[60] (58)
3M	270-330	155-215 or 385-445	[60] (68)	40-100 or 500-560	[70] (75)	75-15* or 675-615	[70] (75)	190-130 or 730-790	[65] (62)
4M	360-440	245-325 or 475-555	[70] (71)	130-110 or 590-670	[70] --	15-95* or 705-785	[70] (69)	100-20* or 820-900	[70] (68)
5M	450-550	335-435 or 565-665	[70] (81)	220-320 or 680-780	[70] --	105-225 or 795-895	[70] --	10-90* or 910-1010	[70] (82)

kHz band a simple low pass filter can provide sufficient rejection of the upper sideband.

To minimize the phase distortion in the pass band, it was decided to use a Bessel function filter to perform the low pass requirements for the transmit path. The inclusion of a pilot tone in the final configuration modified this decision somewhat since the pilot tone output as derived by the divider chain is a squarewave and requires filtering to remove the higher harmonics. Thus a combination filter was designed using a Bessel function section and a Butterworth section in tandem. The Bessel function section provides the proper attenuation of the out of band products with minimum phase distortion of the 5 to 25 kHz band while the Butterworth section with a high cutoff frequency provides the additional attenuation to filter the pilot tone harmonics. This filter was designed, built and tested and its response measurements are shown in Table 2-2.

TABLE 2-2.
MEASURED LOW PASS FILTER CHARACTERISTICS

f kHz	Attenuation (db)	Envelope Delay (Microseconds)
5.0	0.2	-1
10.0	0.2	0
15.0	0.5	+1
20.0	0.8	+1
25.0	1.5	+3
30.0	3.3	+5
60.0	18.5	+ 55
90.0	45.0	---
120.0	51.3	---
150.0	62.5	---
200.0	65.6	---
250.0	66.7	---

- RECEIVE FILTER REQUIREMENTS

Using a test set-up similar to that shown in Figure 2-1, the upconversion or receive portion modulation products were measured. These results are tabulated in Table 2-3.

From Table 2-3 it is seen that the undesired inband products (i.e., those in the 90 to 110 kHz band) are at least 60 dB below the input signal (see the asterisk values). Using the same criteria that out-of-band distortion products be suppressed to the level of undesired inband products as a minimum then the out-of-band rejection required is 60 dB from 120 to 140 kHz with no returns higher than 50 dB up to 370 kHz. This steep requirement on the receive filter could be met with a Cauer type LC design but for the purpose of the breadboard feasibility tests conducted under this effort a readily available

TABLE 2-3.

Modulation Products (Up Conversion)

M = Modulating Frequency = 5-25 kHz

C = Carrier Frequency = 115 kHz

		C= 115 kHz		2C= 230 kHz		3C= 345 kHz	
	f kHz	C± f kHz	C= 115 kHz dB	2C± f kHz	C= 230 kHz dB	3C± f kHz	C= 345 kHz dB
M	5-25	110-90 120-140	[0] (0)	225-205 235-255	[40] (38)	340-320 350-370	[10] (11)
2M	10-50	105-65* 125-165	[60] (62)	220-180 240-280	[60] (67)	335-295 355-395	[60] (57)
3M	15-75	100-45* 130-90	[60] (67)	215-155 245-305	[70] (72)	330-275 360-420	[70] ---
4M	20-100	95-15* 135-216	[70] (72)	210-130 250-330	[70] (84)	325-245 365-445	[70] ---
5M	25-125	90-10 140-240	[70] (79)	205-105* 255-355	[70] (82)	320-220 370-470	[70] ---

filter with a cut off, at 108 kHz was used. This filter provides >40 dB attenuation at 120 kHz and >70 dB at 132 kHz with >50 dB attenuation from 160 kHz to beyond 600 kHz. The inband characteristics of the filter are less than 1.5 dB attenuation spread from 90 to 108 kHz with 15 microseconds envelope delay over the same band. In the final configuration a filter with these characteristics should be adequate.

• BACK TO BACK TESTS

Using the transmit and receive filters and modulators, the two were tied in tandem for back-to-back tests. The test set-up used is shown in Figure 2-3. The input signal used was a LORAN C signal received from a 9-foot whip antenna utilizing the LORETS receiver front end. An oscilloscope comparison of input signals (point A) versus output signal (point B) was made to determine the effect of the dual frequency translation process. The data obtained from this test was visual and the observations revealed very little pulse distortion between the two sets of signals.

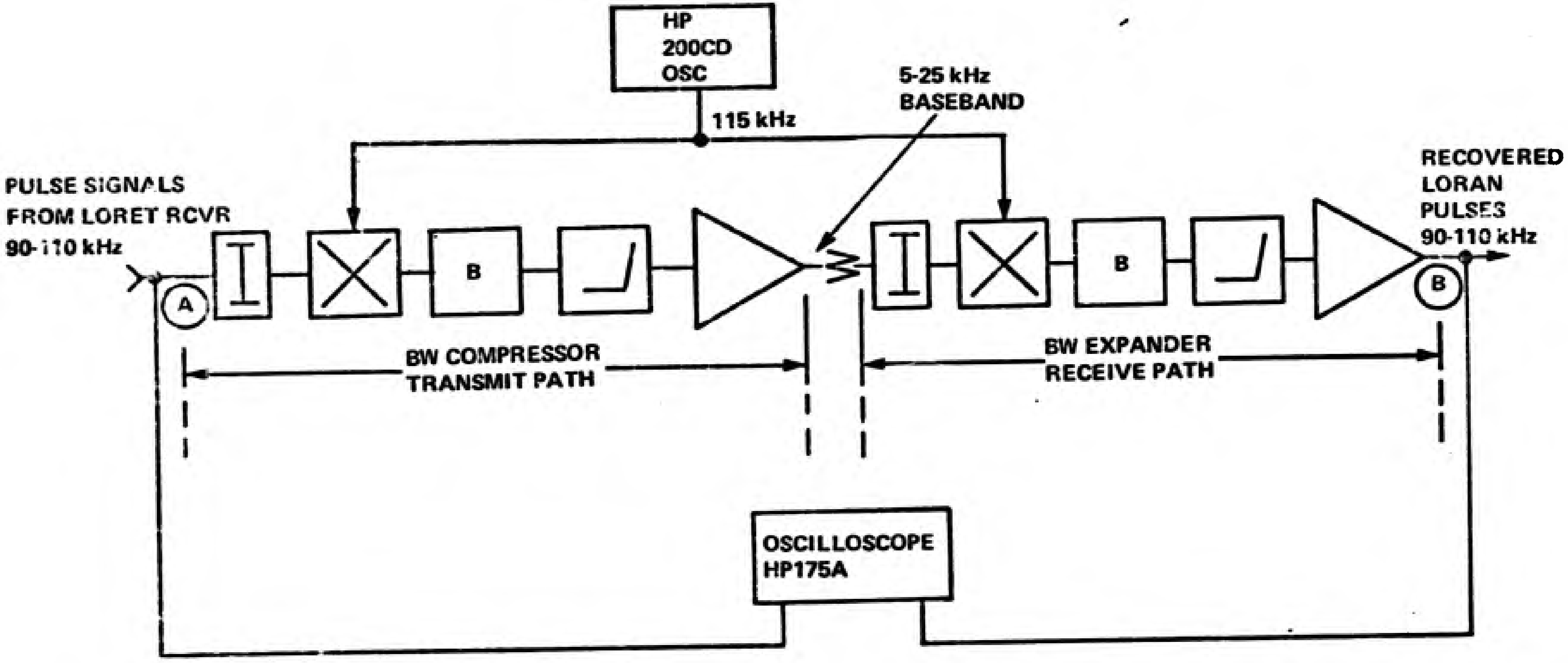


FIGURE 2-3. BACK-TO-BACK SETUP

Additional tests were conducted using an oscillator as the input in place of the LORAN C signal from the air with the output measured using a selective VTVM. Under these conditions the insertion loss spread over the frequency range was less than 2.0 dB from 90 to 110 kHz. Attempts to determine the envelope delay were not conclusive due to malfunctioning of the test equipment used (Wandel & Goltermann LD2) but indicated approximately 5 microseconds from 90 to 100 kHz and approximately 25 microseconds from 110 to 110 kHz which do not agree with earlier measurements of 15 microseconds on the individual filters.

2.2 IMPLEMENTATION OF THE CARRIER SYNCHRONIZATION

The basic scheme for deriving the receive portion carrier employs a phase-locked-loop controlled by the transmitted pilot tone at 1/4 of the oscillator frequency. The output of the VCO is divided by four and applied to one input of a phase comparator and the pilot tone is applied to the other input. The resulting DC error voltage, after filtering, is used to control the bias on a constant current charging source for the timing capacitor of the VCO. Several packaged VCO's and phase lock loops are commercially available which require only a few external components to adjust the free running frequency of the VCO and the bandwidth of the feedback path. A typical one is the Signetics SE565 which contains a VCO, a phase comparator, a low pass filter and a buffer amplifier. The VCO requires two external components, a resistor and a capacitor to adjust the free running frequency and the low pass filter requires a capacitor to adjust the break point of the feedback loop response. The inclusion of a divider between the VCO output and the comparator input enables the comparison to be made at 1/4 of the operating frequency. A circuit diagram of the basic carrier generation recovery is shown in Figure 2-4 and the calculations for the operating parameters are given in Appendix B.

The bandwidth of the feedback loop was set at approximately ±500 Hz by using a 15 uF capacitor for C2 and the tracking range was reduced to a minimum by

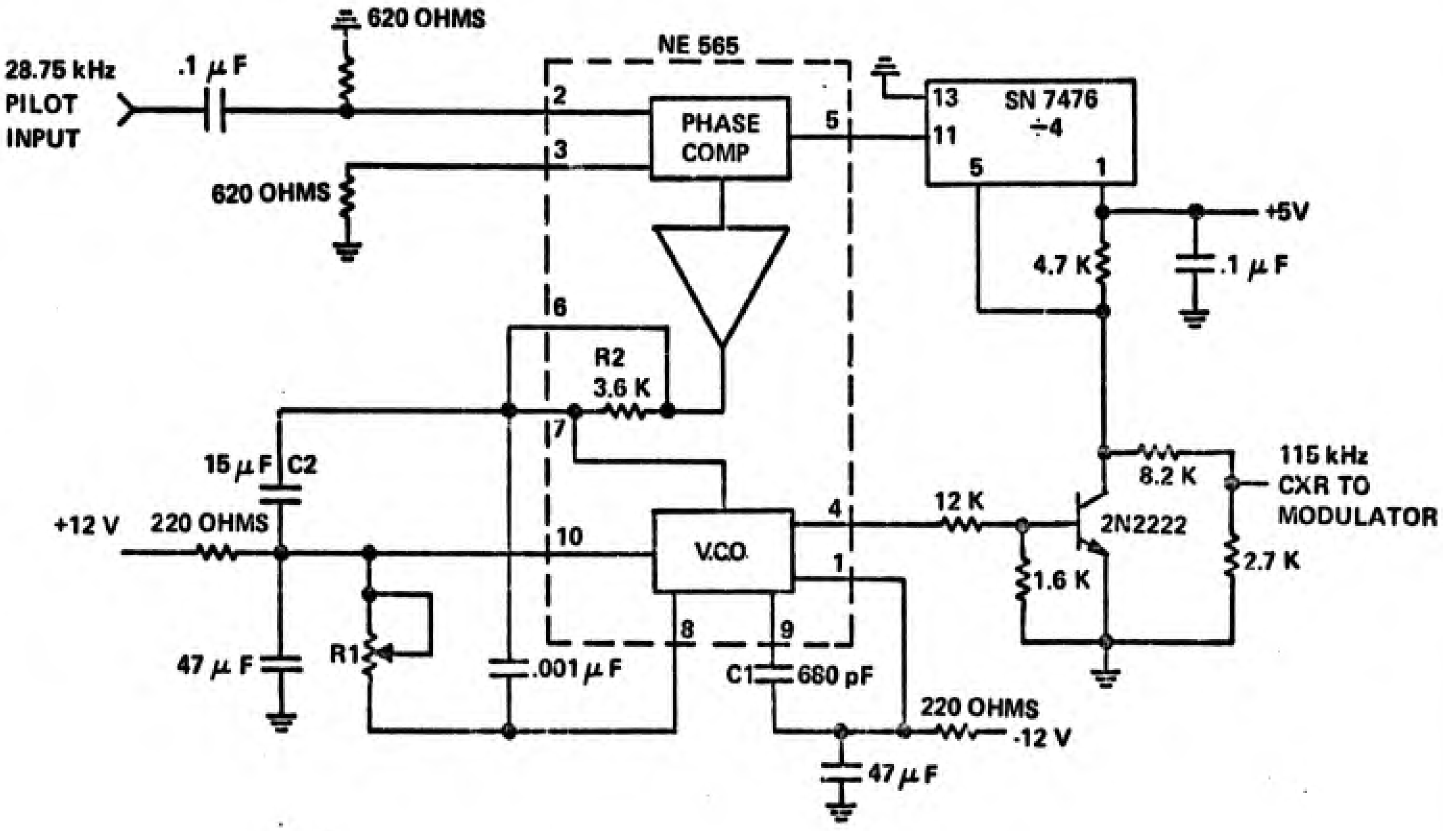


FIGURE 2-4. CARRIER GENERATION

keeping the gain of the amplifier low. Parameters given in the data sheet for the NE565 show the minimum tracking range for the unit operating on a ±12 volt supplies to be approximately ±10 percent of the free running frequency. This was checked using the test setup shown in Figure 2-5. The free running frequency of the VCO was adjusted to 115 kHz, a sine wave generator at a level of 50 mv RMS was used for the input signal. The output of the VCO and the input signal were displayed on an oscilloscope to determine when the signals were in lock and the capture points and tracking range was measured using a frequency counter to record the frequency of the input signal. The data obtained in this test is shown below:

VCO Free Running Frequency = 114.75 kHz
Low Side Capture = 28.53 x 4 = 114.12 kHz
High Side Capture = 28.77 x 4 = 115.08 kHz
Capture Range = 0.95 kHz or ±480 Hz
Low Side Tracking Limit = 26.353 x 4 = 105.412
High Side Tracking Limit = 30.935 x 4 = 123.740
Tracking Range = 18.328 kHz or ±9.164 kHz

Attempts to determine the effects of adjacent or interfering tones on the carrier synchronization using an additional oscillator input in parallel with HP200CD of Figure 2-5 showed that the loop was very stable., once it was in lock, to discrete tones at levels in excess of the 50 mv rms desired signal level to within ±2 kHz of the controlling frequency. However, at discrete intervals of approximately 400 Hz signs of jitter were evident. Improving the power supply decoupling reduced the magnitude of the jitter but did not completely eliminate it. This is a problem area and needs careful attention in laying out the final units, particularly the grounding of the oscillator with respect to the divider package. In the present breadboard configuration, separate power supplies are used since the VCO operates from ±12V supplies and the divider from a + 5V logic supply. Also, the inclusion of an extra ground return for the interfering signal adds yet another ground loop which

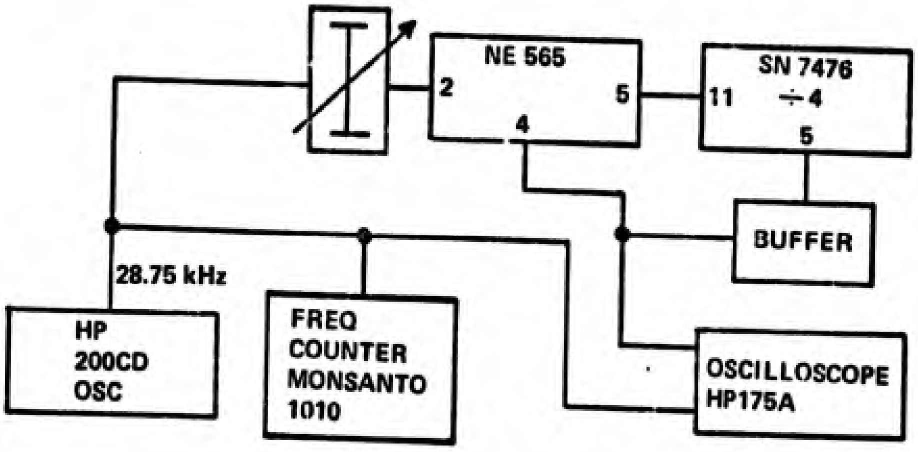


FIGURE 2-5. TRACKING RANGE TEST

only aggravates the situation. It is anticipated that the problem will be resolved in the repackaging of the breadboard to a final configuration where ground loops and power supply problems will be minimized. It was not expedient to expend additional effort to this problem since feasibility was the primary goal and final implementation was of secondary nature.

2.3 IMPLEMENTATION OF THE BREADBOARD SYSTEM FOR BW REDUCTION DEMONSTRATION

For the breadboard system feasibility demonstration, both ends of the path were assembled using the circuits previously discussed. The transmit end contained a carrier source, modulator and the low pass filter described earlier. The receive end contained the carrier recovery circuit, modulator and filter as described above.

For the transmit end the carrier source consisted of a VCO and phase-lock-loop controlled by an external HP200CD oscillator. The phase-lock-loop operated at 1/4 the free running frequency of the VCO as at the receive end. The VCO provides a square wave carrier to the modulator at approximately 300 mv peak to peak level to minimize carrier feed through and unwanted distortion products. The divided by four output from the VCO is used to provide the pilot tone as well as the comparison input to the phase-lock-loop. The pilot tone is summed with the output from the modulator into the active low pass filter and a level adjustment is provided to set the pilot tone to approximately 50 mv RMS. (In the final repackaged configuration the VCO will be replaced with a crystal controlled oscillator.) The combined output from the low pass filter is amplified to the required 3 V peak to peak amplitude (of the LORAN pulses) to drive the LORETS RF transmitter. A diagram of the transmit path is shown in Figure 2-6.

The receive end of the system is shown in Figure 2-7 and contains the carrier recovery circuit, modulator and low pass filter. The output of the low pass filter is amplified and fed to a bandpass filter to reduce the level of the modulated pilot tone which appears at 86.25 kHz. This unwanted pilot disturbs the LORAN Receiver (ADL-21M) since it is at a level greater than 10 percent of the peak amplitude of the weaker slave station that is received. With the inclusion of the amplitude compression technique, the effect of this pilot tone would be far less. The impact of the bandpass filter on the final output is to introduce a small amount of both phase and amplitude distortion. The response of this filter in the passband shows an amplitude response of -1.0 dB from 91 to 109 kHz and an envelope delay of 35 microseconds over the same frequency range. However, the envelope delay from 94 kHz to 106 kHz (±6kHz) is less than 10 microseconds and its degradation of the system time interval response is difficult to assess under the present low signal level conditions.

The system was setup as shown in Figure 2-8 and data taken on the ADL-21M to demonstrate the effects of the bandwidth compression on the overall

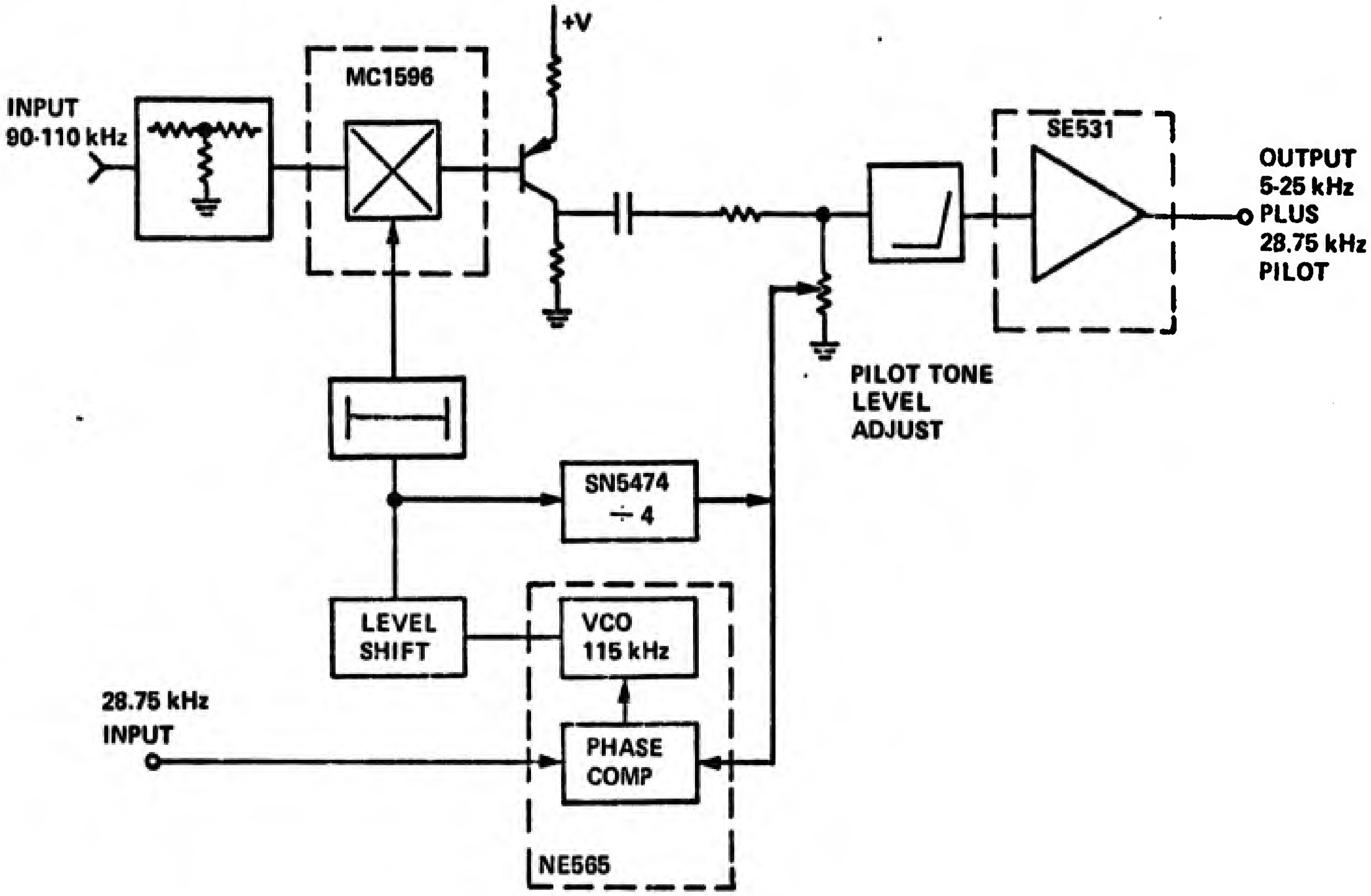


FIGURE 2-6. TRANSMIT PATH DIAGRAM

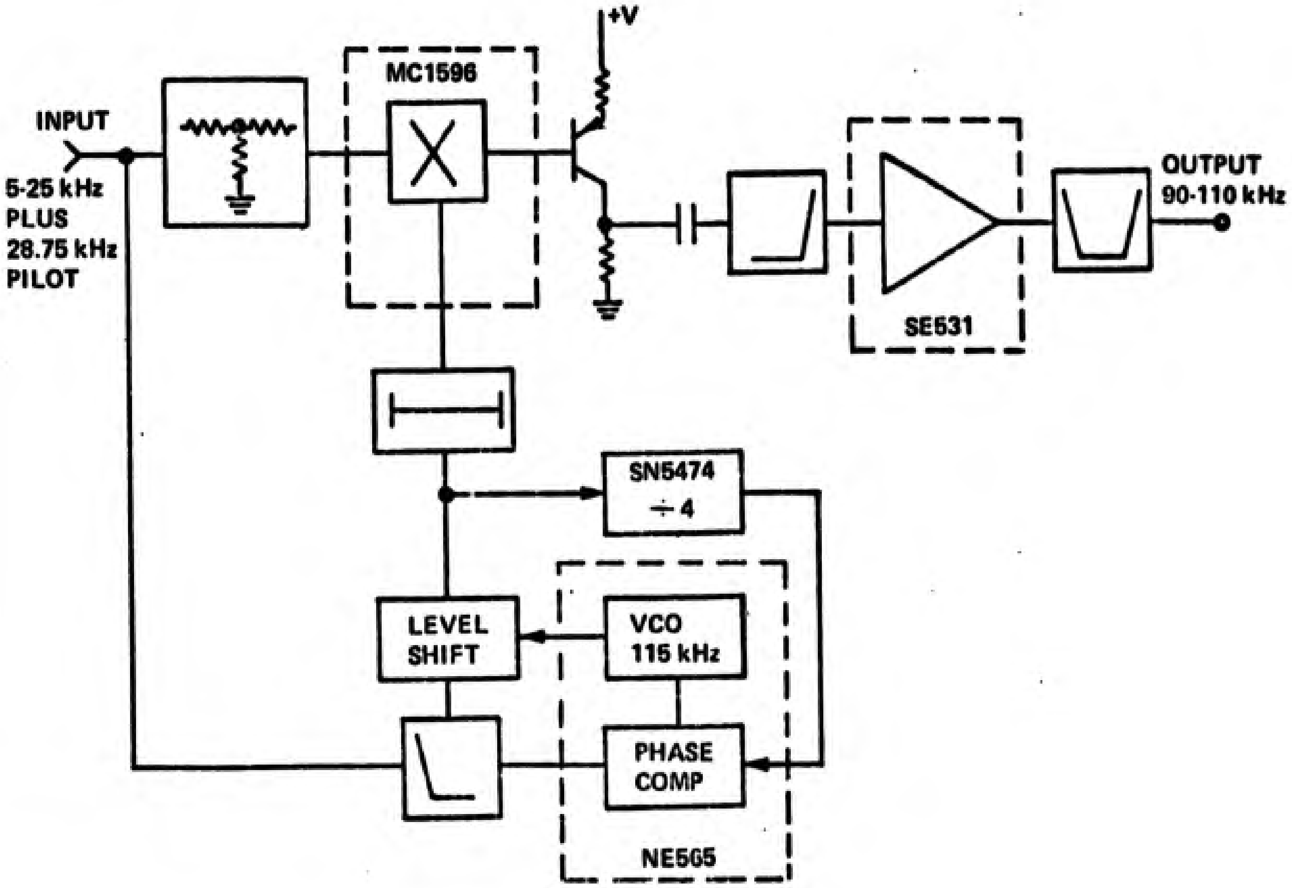


FIGURE 2-7. RECEIVE PATH DIAGRAM

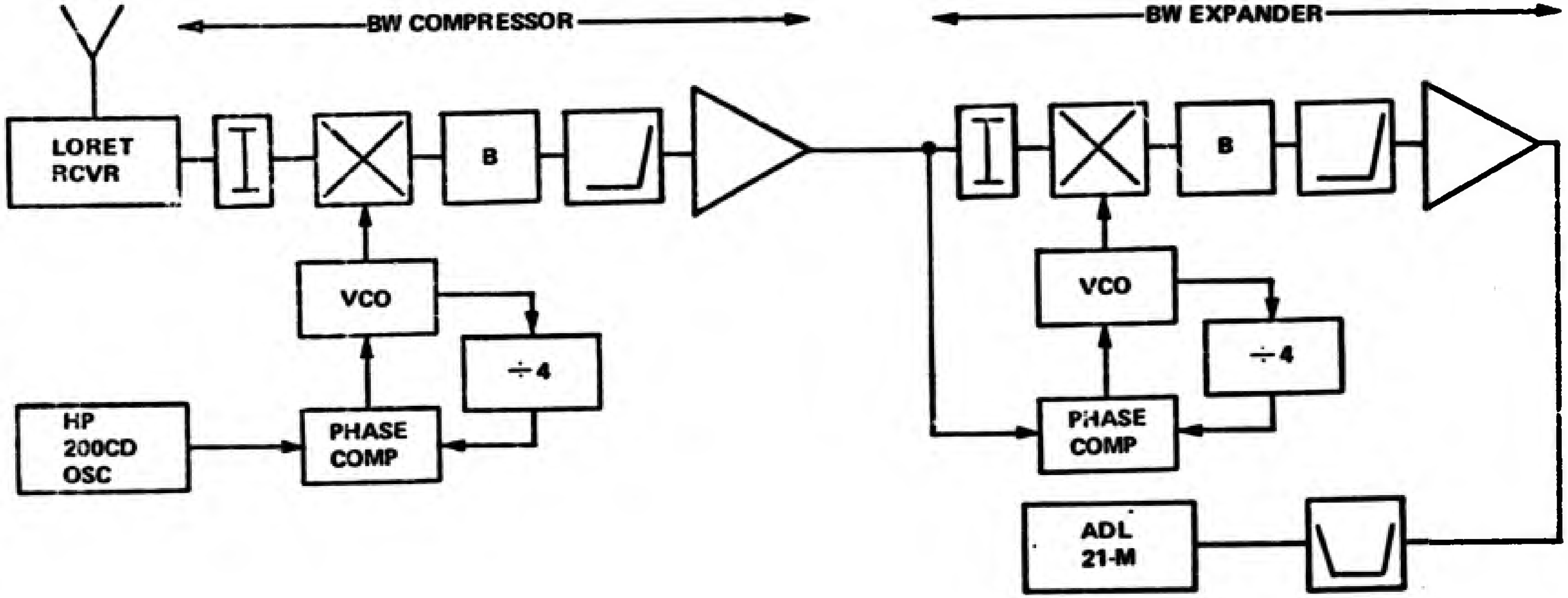


FIGURE 2-8. TEST SETUP FOR BW REDUCTION TIME INTERVAL MEASUREMENTS

retransmission system. Two sets of data were taken, one on the receiving system which included both the bandwidth compression and subsequent expansion, the other set used only the LORAN receiving antenna with the ADL-21M.

From this data it can be seen that some distortion is introduced by the bandwidth compressor which is small for the strong slave station using either 3rd cycle or 5th cycle indexing. On the weak slave the distortion is a minimum using the 5th cycle index and more noticeable using the 3rd cycle index.

TEST DATA

TDA = CAPE FEAR/JUPITER
TDB = CAPE FEAR/DANA

THIRD CYCLE INDEXING

	With BW Compressor	Direct Antenna
TDA (Max.)	11819.6 usecs.	11819.5 usecs
TDA (Min.)	11819.1 usecs.	11819.3 usecs
*TDA (Average)	11819.35 usecs	11819.4 usecs
TDB (Max.)	70480.6 usecs	70479.3 usecs
TDB (Min.)	70478.0 usecs	70478.0 usecs
*TDB (Average)	70479.3 usecs	70478.65 usecs

For the Direct Antenna TDB most repeated = 70478.8 usecs.

For the BW Compressor no preferred value.

Note: The above range of readings were taken over a period of several hours with the ADL-21M display indicating TDA and TDB alternately. The reception of slave B is unreliable due to lack of signal strength for optimum AGC and Index Readings using the 3rd cycle index.

*These readings are calculated from the data taken. The readings were repeated using the 5th cycle for indexing with the following results.

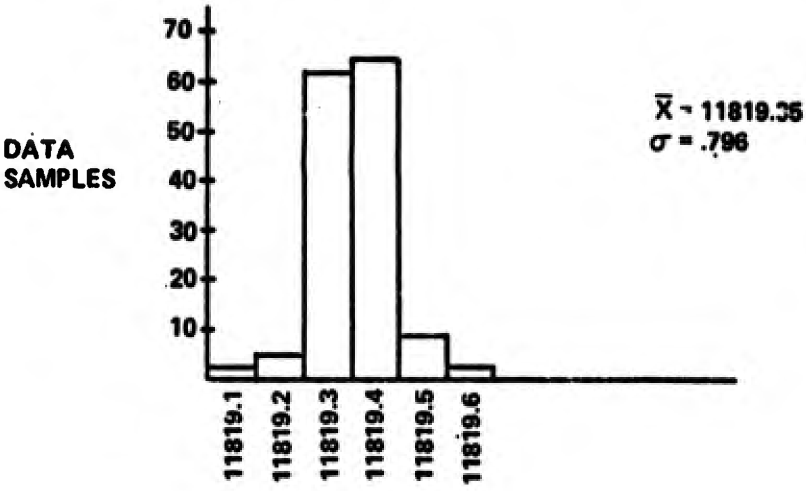
FIFTH CYCLE INDEXING

	With BW Compressor	Direct Antenna
TDA (Max.)	11819.3 usecs	11819.5 usecs
TDA (Min.)	11819.2 usecs	11819.4 usecs
*TDA (Average)	11819.25 usecs	11819.45 usecs
TDB (Max.)	70479.5 usecs	70479.4 usecs
TDB (Min.)	70479.2 usecs	70479.3 usecs
*TDB (Average)	70479.35 usecs	70479.35 usecs

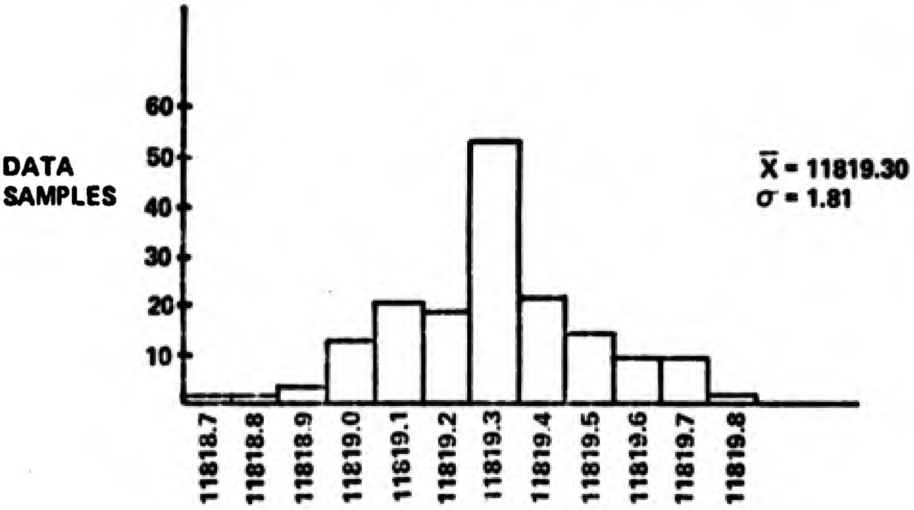
A second set of data was taken to quantitatively assess the effect of the Bandwidth Reduction Technique. This data is plotted in Figures 2-9 and 2-10 illustrating the readings obtained (i.e., TDB and TDB) from the ADL-21 LORAN receiver with and without the Bandwidth Compressor. Due to time constraints in the program it was not possible to take extensive sets of data, however, the data illustrated is typical and serves as a vehicle for quantitatively assessing the overall system performance. The calculations that were made on this data appear on the data sheets of Appendix C.

Figure 2-9 illustrates the results of TDA (Cape Fear and Jupiter) for the case with the BW Compressor in and out of the circuit. Note that the addition of the BW Compressor in this case results in a slight shift of the mean value (.05 microseconds) which corresponds to an error of about 47 feet. The spread in the data increases from $\sigma = .796$ to 1.81 using the BW Compressor.

Figure 2-10 illustrates the results of TDB (Cape Fear and Dana). In this case the shift in the mean value with the addition of the BW Compressor was equal to .21 microseconds corresponding to an error of 211 feet. The spread in the data increased from $\sigma = 2.06$ to 6.2. The increase in σ was of the same order for TDB as for TDA. The greater shift in the mean value of TDB is due to the lower input S/N ratio due to the remoteness of St. Petersburg to Dana. There are many times during the day when Dana cannot be received at all, therefore conclusions on these statistics should keep this factor in mind.



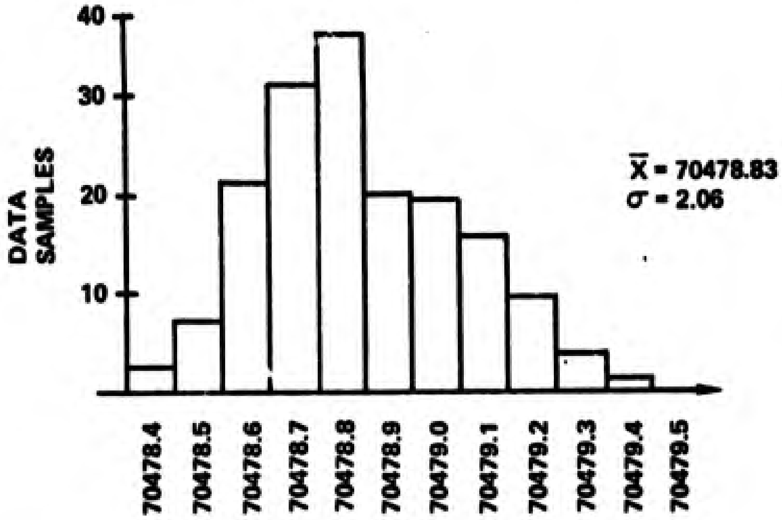
(A) DIRECT – TDA



(B) THROUGH BW COMPRESSOR – TDA

FIGURE 2-9. TDA STATISTICS

(A) DIRECT – TDB



(B) THROUGH BW COMPRESSOR – TDB

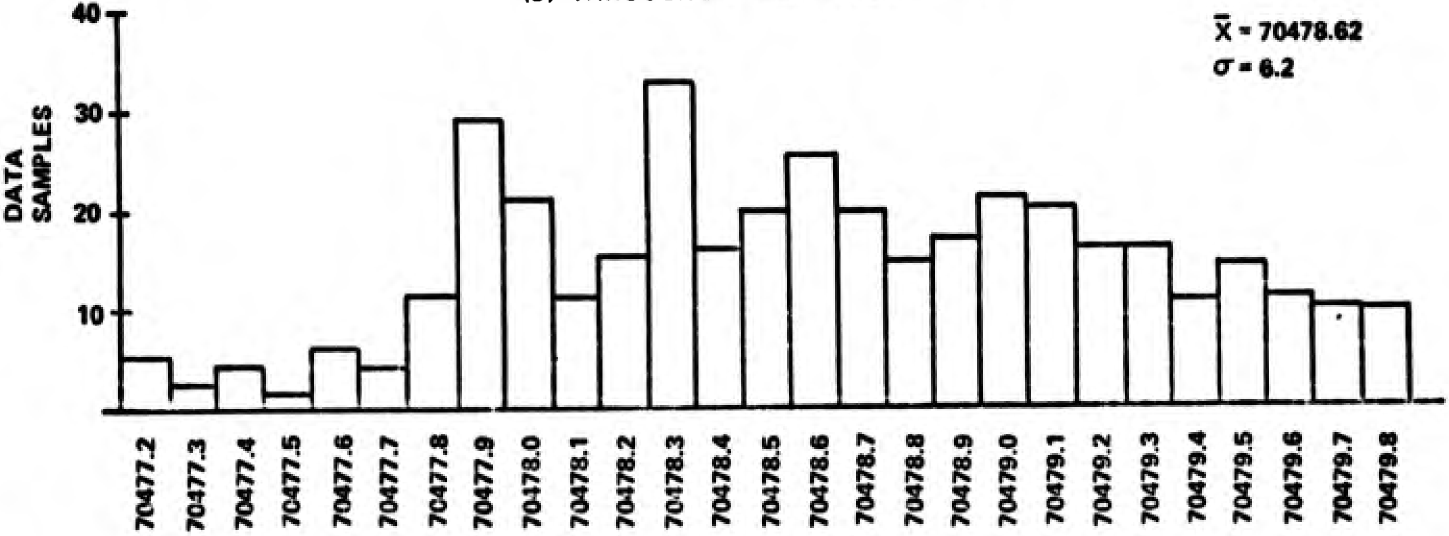


FIGURE 2-10. TDB STATISTICS

It is therefore clear from the data that under a good S/N condition the BW Compressor adds little degradation to the total system performance. Under poor S/N conditions the degradation is somewhat more noticeable. The reasons for this can be seen by a more detailed appraisal of the hardware involved in the BW expander portion of the system. The low pass filter used in the receiver has a cut-off at 108 kHz instead of the desired 110 kHz which reduces the effective receive bandwidth. In the final configuration the dynamic range of the signals will be greatly reduced by the inclusion of the amplitude compressor in the transmit portion. Thus, for the weaker stations, the ratio of the wanted signal to the pilot tone will be much greater, which will result in a better signal to unwanted translated pilot tone ratio at the output. The presence of this "pilot" tone at 86.25 kHz at a high level disturbs the operation of the AGC circuits in the ADL-21M and results in an apparent poor signal to noise ratio. The attempts in the breadboard system to filter out this tone with the additional bandpass filter restricts the bandwidth of the output pulse train to less than 90 to 110 kHz which also effectively lowers the signal-to-noise ratio of the stronger signals by removing power from the signal bandwidth. The type of filter used has more envelope delay within the passband than the filter which will be used in the final configuration. This also adds to the distortion of the present retranslated output pulse stream.

The data taken shows that the variance of the weaker slave TDB is far greater than that of the stronger slave (TDA) when the readings using the direct antenna are analyzed (comparing Figures 2-9a. and 2-10a). This indicates a poorer signal to noise ration for the weaker slave. A comparison of the variances of the two slaves when using the BW compressor against those through the direct antenna indicates the weaker slave undergoes the same percentage distortion as the stronger slave. The greater shift of the mean value of the weaker slave with the BW compressor also illustrates the effective lowering of the signal to noise ratio due to the interfering pilot tone. The attempts to remove this tone by including a filter which restricts the passband width has succeeded in degrading the system performance of the stronger slave without providing sufficient rejection when the weaker slave is being measured. This suggests that the effect of restricting the signal bandwidth contributes more to the distortion than the interference caused by the lower signal to pilot tone levels on the AGC circuits within the ADL 21-M. Thus the use of the correct filter bandwidth for the receive path and the correct pilot tone to signal levels will reduce this spreading of the variances considerably in the final repackaged unit.

2.4 REPACKAGING APPROACH

In the repackaged system the transmit portion (the bandwidth compressor) and the receive portion (the bandwidth expander) will be packaged as separate entities. The compressor forms a part of the LORET package while the expander will become part of the AN/ARC-34 modification.

The compressor will be packaged in a shielded compartment with its own power converter operating from a 16 volt battery. The local oscillator for the carrier frequency will be a TCXO with two frequency outputs for the modulator and one for the pilot frequency. The modulator and pilot combining buffer will be built as part of the low pass filter to reduce the high impedance interface points to a minimum. The final output amplifier will be well isolated from the power supply to reduce unwanted feedback paths to the modulator both by radiation and conduction mechanisms. Controls will be provided for pilot tone level and frequency adjustment over a limited range.

The expander will be packaged as an add-on unit to the AN/ARC-34. The carrier synchronization circuitry will be in one shielded compartment and the modulator and bandpass filter with the output amplifier in another. Isolation between the individual units and the common power source will be provided by means of decoupling on each unit. Since the bandpass input characteristics of the new LORAN receivers are unknown at present, the filter in the compressor will be adequate to permit operation with the existing ADL-21M although some relaxation of the stop-band rejection may be permissible with the newer generation of receivers.

The jitter problems encountered in the breadboard phase which were attributed to coupling via power supply and ground leads and pick-up of radiated signals from local interference sources will be nullified by using shielded compartments and a single power source for each of the two units. The isolation between the individual submodules and the common power source will also be improved in the repackaging which will also help to remove inter-reaction problems.

SECTION 3
CONCLUSIONS

It is shown in Appendix A that a system bandwidth reduction system such as the one implemented under this contract theoretically transmits the LORAN signal with no distortion since the local oscillator used for upconversion is identical in phase with the conversion local oscillator. Any corruption that the signal experiences in this process is due to the following peripheral sources:

1. Phase error in the pilot tone recovery phase lock loop
2. Phase response distortion in the final bandpass filter of the receiver.

The first source of distortion, phase lock loop error, has been made arbitrarily small by the use of stable oscillators for the loop oscillator and an arbitrarily narrow loop filter.

The second source of distortion, bandpass filter phase response, is perhaps the most critical. This filter has the task of separating the desired signal from the unwanted sideband of the upconversion process. As shown in the data taken on the breadboard model of the BW Compressor this filter must be optimized to reject the translated pilot tone and yet not add an appreciable amount of envelope delay to the system.

From the results of the bench tests performed and described in Section 2 and the statistical analysis of the data taken it can be concluded that the use of a BW Compressor as designed and implemented herein is indeed feasible provided that the system incorporates the following end-to-end parameters:

1. That the out-of-band rejection of undesired sidebands in the down conversion and upconversion be kept to 40 dB minimum. The final system design goal will be to provide 60 dB rejection of these sidebands.

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2. That the total in-band envelope delay be kept to less than 15 microseconds in the final bandpass filter. It is suspected that the 35 microseconds envelope delay of the filter used in the breadboard BW Compressor contributed to the slight shift in the TDA and TDB mean value readings and hence caused a slight error. This error can be minimized by designing the final filter with 15 microseconds envelope delay.

3. That the total retransmission bandwidth be kept to 2 x 28.75 kHz = 57.5 kHz at the 3 dB points. This results in a net savings of required retransmission bandwidth from >220 kHz to 57.5 kHz or about a 4 to 1 reduction.

Based on the results and analysis and demonstrated feasibility it is recommended that the BW Compressor be repackaged into the Slice #3 configuration and that the AN/ARC-34 be modified with the BW Expander circuitry to permit further evaluation of this concept.

APPENDIX A

PHASE ERROR IN FREQUENCY TRANSLATION PROCESSES

It is desired to determine the phase error that occurs when an arbitrary waveform is shifted in frequency by heterodyning and subsequently returned to its original frequency.

Assume the input signal is of the form

$$f_1(t) = f(t) \cos \omega_o t$$

where $\frac{\omega_o}{2\pi} \cong 100$ kHz and f(t) is a baseband waveform, in this case, given by the function $f(t) = (kt)^2 \, e^{-2(kt-1)}$.

Mix f_1 with a local oscillator signal

$$f_2(t) = \cos (\omega_L t + \theta_L)$$

where ω_L is the local frequency, $\omega_L > \omega_o + B$, and θ_L is its arbitrary phase.

The resulting signal is given by

$$f_3(t) = f(t) \cos \omega_o t \cos (\omega_L t + \theta_L)$$

$$= f(t) \cos \left(\left[\omega_L - \omega_o\right] t + \theta_L\right) + f(t) \cos \left(\left[\omega_L + \omega_o\right] t + \theta\right)$$

where the gain factor has been neglected. (For high level constant amplitude of the mixing signal $f_2(t)$ the amplitude of $f_3(t)$ will be directly proportional to the amplitude of $f_1(t)$.)

This signal is now lowpass filtered to remove the $\cos (\omega_L + \omega_o)t$ term, leaving the term,

$$f_4 = f(t) \cos \left(\left[\omega_L - \omega_o\right] t + \theta_L\right)$$

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To aid phase-coherent recovery, a pilot tone is added, but at a reduced frequency to preserve bandwidth. This is generated by frequency division of the local oscillator.

$$f_5(t) = \cos\left(\frac{\omega_L t + \theta_2}{n}\right)$$

$$= \cos\left(\frac{\omega_L}{n} t + \frac{\theta_2}{n}\right)$$

The composite transmitted signal then is

$$f_6(t) = f(t) \cos\left(\left[\omega_L - \omega_o\right] t + \theta_L\right) + \cos\left(\frac{\omega_L}{n} t + \frac{\theta_L}{n}\right).$$

In the transmission process, there may be some phase distortion of the pilot with respect to the modulation since it is at the band edge of the receiver. The received signal will be

$$f_7(t) = f(t) \cos\left(\left[\omega_L - \omega_o\right] t + \theta_L + \theta_T\right) + \mathrm{Cos}\left(\frac{\omega_L}{n} t + \frac{\theta_L}{n} + \theta'_T\right)$$

where θ_T and θ'_T are the transmission delays.

From f_7, the pilot tone is recovered by a phase locked loop. The output of the loop VCO is shifted 1/4 cycle from the pilot tone phase which is correctable by a fixed delay in the carrier recovery system. There will be a small phase error if there is a frequency difference between the incoming pilot signal and the natural frequency of the VCO. This error is

$$T_e = \frac{1}{4f} - \frac{1}{4f'}$$

or $$\theta_{CRL} = \frac{1}{4} \frac{2\pi}{T'} \left(\frac{1}{f} - \frac{1}{f'}\right) \text{ where } T' = \frac{1}{f'}$$

$$= \frac{\pi}{2} f' \left(\frac{f' - f}{ff'}\right) = \frac{\pi}{2} \frac{\Delta f}{f} = \frac{\pi}{2} \frac{\Delta\omega}{\omega}$$

which will also be taken to include any time error inaccuracy in the delay and the loop.

The recovered pilot, then, after shifting is

$$f_8(t) = \cos(\omega_L t + \theta_L + \theta_T + \theta_{CRL})$$

The received signal f_7, after lowpass filtering to remove the pilot tone is

$$f_9(t) = f(t) \cos\left[(\omega_L - \omega_o) t + \theta_L + \theta_T\right]$$

Mixing f_8 and f_9 gives

$$f_{10}(t) = f(t) \cos\left[\omega_o t + \theta_T' - \theta_T + \theta_{CRL}\right] + f(t) \cos\left[(2\omega_L - \omega_o)t + 2\theta_L + \theta_T' + \theta_T + \theta_{CRL}\right]$$

where again gain terms have been ignored.

Using a lowpass filter to remove the term at $2\omega_L - \omega_o$ (which must be quite selective) gives

$$f_{11}(t) = f(t) \cos\left[\omega_o t + (\theta_T' - \theta_T) + \theta_{CRL}\right]$$

This resultant signal, $f_{11}(t)$, is identical to the original waveform $f_1(t)$ except for the phase errors due to the differential phase error in the baseband, $(\theta_T' - \theta_T)$ and the carrie recovery loop phase error θ_{CRL}. In an actual system, the part of $(\theta_T' - \theta_T)$ error which is due to transmitter-receiver characteristics can be compensated in the delay at the output of the carrier recovery. Proper loop design can keep θ_{CRL} small, (since it is proportional to the absolute frequency error of the phase lock loop VCO as a percentage of the oscillator frequency) so the only significant error is due to the $(\theta_T' - \theta_T)$ error caused by the transmission medium.

APPENDIX B

PHASE-LOCK-LOOP CALCULATIONS

From the data sheet for the NE-565A the VCO free running frequency f_o is given by

$$f_o = \frac{1}{4R1C1}$$

for f_o = 115 kHz and R1 = 3.3 KΩ

$$C1 = \frac{1}{460 \times 10^3 \times 33 \times 10^3} = \frac{10^{-9}}{1.52} = 680 \text{ pf}$$

Lock Range f_L is given by

$$f_L = \pm \frac{8 f_o}{VCC}$$ where VCC is total DC voltage applied between pins 1 and 10.

$$= \pm \frac{920}{24} = \pm 38.3 \text{ kHz}$$

$$\text{or} \quad \pm \frac{38.3}{115} \times 100\% = 33.3\%$$

To restrict the lock range to approximately ±10% the gain setting resistor is reduced to zero ohms.

Capture Range f_c is given by

$$f_c = \frac{1}{2\pi}\sqrt{\frac{2\pi f_L}{\tau}}$$

Where τ = R2C2 and R2 = 3.6 KΩ (an internal resistor).

In the system C2 is 15 uF in the Receiver

$$\therefore \tau = 15 \times 10^{-6} \times 3.6 \times 10^{-3}$$

$$= 54 \times 10^{-3}$$

$$\therefore f_c = \frac{1}{6.28} \sqrt{\frac{6.28 \times 38.3 \times 10^3}{54 \times 10^{-3}}}$$

$$= \frac{1}{6.28} \; 2.11 \,.\, 10^3$$

$$= .336 \text{ kHz}$$

By reducing the lock range to approximately 10% of f_o f_c reduces to

$$\frac{1}{6.28} \sqrt{\frac{6.28 \times 11.5 \times 10^3}{54 \times 10^{-3}}}$$

$$= \frac{1}{6.28} \; 1.152 \; 10^3$$

$$= .184 \text{ kHz}$$

With C2 selected as 6.8 μF in the Transmitter.

τ reduces to $6.8 \times 10^{-6} \times 3.6 \times 10^3$

$= 24.5 \times 10^{-3}$

then f_c (with f_c at ±10% f_o)

$$= \frac{1}{6.28} \sqrt{\frac{6.28 \times 11.5 \times 10^3}{24.5 \times 10^{-3}}}$$

$$= \frac{1}{6.28} \; 1.72 \,.\, 10^3$$

$$= 0.274 \text{ kHz}$$

APPENDIX C

DATA SHEETS AND CALCULATIONS TDA AND TDB

DW COMPRESSION - TDB

Time Interval (Microseconds)	n_x	X	$n_x X$	$n_x (X-\bar{X})^2$		
70477.2	5	1	5	$5 (14.2)^2$	=	1008
70477.3	2	2	4	$2 (13.2)^2$	=	348
70477.4	4	3	12	$4 (12.2)^2$	=	595
70477.5	2	4	8	$2 (11.2)^2$	=	250
70477.6	6	5	30	$6 (10.2)^2$	=	624
70477.7	4	6	24	$4 (9.2)^2$	=	338
70477.8	13	7	91	$13 (8.2)^2$	=	874
70477.9	29	8	232	$29 (7.2)^2$	=	1503
70478.0	21	9	189	$21 (6.2)^2$	=	807
70478.1	12	10	120	$12 (5.2)^2$	=	324
70478.2	15	11	165	$15 (4.2)^2$	=	264
70478.3	33	12	396	$33 (3.2)^2$	=	337
70478.4	15	13	195	$15 (2.2)^2$	=	72
70478.5	19	14	266	$19 (1.2)^2$	=	27
$\bar{X}$ 70478.6	25	15	375	$25 (.2)^2$	=	1
70478.7	19	16	304	$19 (.8)^2$	=	12
70478.8	14	17	238	$14 (1.8)^2$	=	45
70478.9	17	18	306	$17 (2.8)^2$	=	133
70479.0	21	19	399	$21 (3.8)^2$	=	303
70479.1	20	20	400	$20 (4.8)^2$	=	460
70479.2	17	21	357	$17 (5.8)^2$	=	571
70479.3	17	22	374	$17 (6.8)^2$	=	786
70479.4	11	23	253	$11 (7.8)^2$	=	669
70479.5	14	24	336	$14 (8.8)^2$	=	1084
70479.6	11	25	275	$11 (9.8)^2$	=	1056
70479.7	10	26	260	$10 (10.8)^2$	=	1166
70479.8	9	27	243	$9 (11.8)^2$	=	1253
	n = 385		5857 =			14910

$\bar{X} = \frac{1}{n} \Sigma X n_x = \frac{5857}{385} = 15.2$

$\bar{X} = 70478.6$

$\sigma^2 = \frac{14910}{385} = 38.72$

$\sigma = 6.2$

DIRECT - TDB

Time Interval (Microseconds)	n_x	X	$n_x X$	$n_x (X-\bar{X})$		
70478.4	3	1	3	$3 (4.3)^2$	=	55.4
70478.5	7	2	14	$7 (3.3)^2$	=	76.2
70478.6	21	3	63	$21 (2.3)^2$	=	111.0
70478.7	32	4	128	$32 (1.3)^2$	=	54.0
70478.8	36	5	180	$36 (.3)^2$	=	3.2
70478.9	19	6	104	$19 (.7)^2$	=	9.3
70479.0	19	7	133	$19 (1.7)^2$	=	54.9
70479.1	16	8	128	$16 (2.7)^2$	=	116.6
70479.2	9	9	81	$9 (3.7)^2$	=	123.0
70479.3	3	10	30	$3 (4.7)^2$	=	66.2
70479.4	1	11	11	$1 (5.7)^2$	=	32.4
	n = 165	165	874			702.2

$\bar{X} = 5.3$

$\bar{X} = 70478.8$

$\sigma^2 = \frac{702.2}{165}$ 4 4.25

$\sigma = 2.06$

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BW COMPRESSION - TDA

	Time Interval (Microseconds)	n_x	X	$n_x X$	$n_x (X-\bar{X})^2$		
	11818.7	1	1	1 x 1 = 1	$1\,(6)^2$	=	36
	11818.8	1	2	2	$1\,(5)^2$	=	25
	11818.9	2	3	8	$2\,(4)^2$	=	32
	11819.0	13	4	52	$13\,(3)^2$	=	117
	11819.1	20	5	100	$20\,(2)^2$	=	80
	11819.2	18	6	108	$18\,(1)^2$	=	18
$\bar{X}$	11819.3	53	7	371	$53\,(0)^2$	=	---
	11819.4	21	8	168	$21\,(1)^2$	=	21
	11819.5	14	9	126	$14\,(2)^2$	=	56
	11819.6	9	10	90	$9\,(3)^2$	=	81
	11819.7	9	11	99	$9\,(4)^2$	=	144
	11819.8	1	12	12	$1\,(5)^2$	=	25
		n = 162		$1135 = \Sigma X n_x$			535 =

$\Sigma (X-\bar{X})^2 \; n_x$

$$\bar{X} = \frac{1}{n} \; \Sigma X n_x = \frac{1135}{162} = 7.0 = 11819.30$$

$$\sigma^2 = \frac{535}{162} = 3.3$$

$$\sigma = 1.81$$

DIRECT - TDA

	Time Interval (Microseconds)	n_x	X	$n_x X$	$n_x (X-\bar{X})^2$		
	11819.1	1	1	1	$1\,(2.56)^2$	=	6.55
	11819.2	4	2	8	$4\,(1.56)^2$	=	76.2
	11819.3	61	3	183	$61\,(.56)^2$	=	19.2
$\bar{X}$	11819.4	65	4	260	$65\,(.46)^2$	=	13.75
	11819.5	8	5	40	$8\,(1.46)^2$	=	17.05
	11819.6	1	6	6	$1\,(2.46)^2$	=	6.05
		140		498	$\Sigma (X-\bar{X})^2 \, n_x = 88.8$		

$$\bar{X} = \frac{1}{n} \; \Sigma X n_x = \frac{498}{140} = 3.56 = 11819.35 \qquad \sigma^2 = \frac{88.8}{140} = .634$$

$$\sigma = .796$$