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Chapter 1

Communications System Box

The communication systems laboratory experiments require the use of several different types of signal processing blocks. In terms of hardware each of these blocks is typically an active electronic circuit, in some cases requiring many components. The communication system box contains eight different electronic subassemblies that are used for generation and detection of modulated signals. The intent of the system box is that its use will allow a reduction in experiment set-up time, thereby giving more time to learn and understand the overall system concepts.

The signal processing blocks incorporated into the system box are the following:

- 2 - analog balanced modulators (multipliers)
- 1 - unity gain level shifter
- 1 - 455KHz I.F. bandpass filter
- 1 - high gain 455KHz I.F. amplifier
- 1 - second-order active lowpass filter with variable cutoff frequency
- 1 - second-order active highpass filter with variable cutoff frequency
- 1 - variable gain summing amplifier with envelope detector output
- 1 - audio amplifier with speaker output

The front panel layout of the system box is shown in Figure 1.1. A brief description of each block including the circuit schematics is given in the following paragraphs.

1.1 Analog Balanced Modulator

The system box contains two analog balanced modulator circuits. Each modulator circuit contains a four quadrant multiplier chip (1496) as shown in the schematic of Figure 1.2. These circuits serve as analog multipliers and thus are used for DSB and AM modulation and demodulation as well as frequency translation in superheterodyne receivers. The $V_c$ input is typically used for the carrier signal. The $V_s$ input is used for the analog message signal during modulation and the
Figure 1.1: Communication System Box Front Panel Layout
1.2 Unity Gain Level Shifter

The unity gain level shifter circuit consists of an op-amp inverting summing amplifier with two inputs. Both inputs are connected in a unity gain configuration. Level shifting is accomplished by connecting one input to a potentiometer that is connected between the +12v and -12v power supply. The schematic for this circuit is shown in Figure 1.3.

The primary use of this circuit is to remove the DC component present on the balanced modulator output. In most applications the DC component is nulled to zero. A case where a non-zero DC component may be desired is in a phase-lock loop (PLL) where a balanced modulator used as a phase detector must drive a voltage controlled oscillator (VCO). Here it is desired to have $V_{dc}$ correspond to the VCO quiescent operating frequency.
1.3 455 KHZ I.F. FILTER

The 455KHz I.F. bandpass filter is a fixed tuned ceramic filter with the input/output terminals brought out to the front panel of the system box. The frequency selectivity of a ceramic filter is due to mechanical vibrations resulting from a piezoelectric effect\(^1\). The equivalent circuit of a basic two terminal ceramic filter resonator is shown in Figure 1.4 along with a typical impedance versus frequency response curve. The resonant frequency, \(f_r\), is given by

\[
f_r = \frac{1}{2\pi \sqrt{L_1 C_1}}
\]  

(1.1)

The so-called anti-resonant frequency, \(f_a\), is given by

\[
f_a = \left[ 2\pi \sqrt{\frac{L_1 C_0}{C_1 + C_0}} \right]^{-1}
\]  

(1.2)

By cascading resonators that are either in a ladder connection or a cascade connection, as shown in Figure 1.5, the desired filter frequency response characteristics can be obtained.

The communication system box uses one of two types of ceramic filters manufactured by Murata Erie North America, Inc. The terminology used to describe the frequency response characteristics of these filters is shown in Figure 1.6. An explanation of each term given in Figure 1.6 is provided in Table 1.1. The type of filter used in the communication system boxes is either an SFU455A or a CFU455D2, where both filters have center frequencies at 455KHz. The detailed specifications for each of these filters are given in Figures 1.7 and 1.8 respectively. Note that the specifications assume either 3K or 1.5K source and load impedances, depending on the filter type. If the impedances are lower than those specified then the center frequency is lowered, the insertion loss increases, and the ripple increases. If on the other hand the impedances are higher than those specified, then the center frequency is higher, with again increased insertion loss and passband ripple. In the experiments these filters will be used to construct a superheterodyne receiver.

\(^1\)Murata Erie North America Inc., Ceramic Filters Catalog No. 59-10, 1985.
1.3. 455 KHz I.F. FILTER

Two-terminal equivalent circuit

Typical frequency response at 455 kHz

Figure 1.4: Two-Terminal Ceramic Filter Resonator

Ladder connection of resonators

Cascade connection of resonators

Figure 1.5: Resonator Connections
Figure 1.6: Ceramic Filter Frequency Response Terminology
Table 1.1: Ceramic Filter Terminology Chart

<table>
<thead>
<tr>
<th>Numbers</th>
<th>Terminology</th>
<th>Symbol</th>
<th>Unit</th>
<th>Explanation of Term</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Center Frequency</td>
<td>$f_o$</td>
<td>Hz</td>
<td>The frequency in the center of the pass-bandwidth. However, the center frequency for some products is expressed as the point where the loss is at its lowest point.</td>
</tr>
<tr>
<td>2</td>
<td>Passband 3dB Bandwidth</td>
<td>BW</td>
<td>Hz</td>
<td>Signifies a difference between the two frequencies where the attenuation becomes 3 dB from the level of the minimum loss point.</td>
</tr>
<tr>
<td>3</td>
<td>Insertion Loss</td>
<td>IL</td>
<td>dB</td>
<td>Expressed as the input/output ratio at the point of minimum loss. (The insertion loss for some products is expresses as the input/output ratio at the center frequency.) Insertion loss = $20\log(E_2/E_1)$ in dB.</td>
</tr>
<tr>
<td>4</td>
<td>Ripple</td>
<td>–</td>
<td>dB</td>
<td>If there are peaks and valleys in the pass-bandwidth, the ripple expresses the difference between the maximum peak and minimum valley.</td>
</tr>
<tr>
<td>5</td>
<td>Attenuation Bandwidth</td>
<td>20 dB</td>
<td>Hz</td>
<td>The bandwidth at a specified level of attenuation. Attenuation may be expresses as the ratio of signal strength to the output signal strength in decibels.</td>
</tr>
<tr>
<td>6</td>
<td>Stopband Attenuation</td>
<td>–</td>
<td>dB</td>
<td>The level of the signal strength at a specified frequency outside the passband.</td>
</tr>
<tr>
<td>7</td>
<td>Spurious Response</td>
<td>SR</td>
<td>dB</td>
<td>The difference in decibels between the insertion loss and the spurious signal in the stop-band.</td>
</tr>
<tr>
<td></td>
<td>Input/Output Impedance</td>
<td>–</td>
<td>Ohm</td>
<td>Internal impedance value of the input and output of the ceramic filter.</td>
</tr>
<tr>
<td></td>
<td>Selectivity</td>
<td>–</td>
<td>dB</td>
<td>The ability of a filter to pass signals of one frequency and reject all others. A highly selective filter has an abrupt transition between a pass-band region and the stopband region. This is expresses as the shape factor—the attenuation bandwidth divided by the pass-bandwidth. The filter becomes more selective as the resultant value approaches one.</td>
</tr>
</tbody>
</table>
SFU 455A

Figure 1.7: SFU455A Specifications

<table>
<thead>
<tr>
<th>Center Frequency (KHz)</th>
<th>3dB Bandwidth (KHz)</th>
<th>Ripple (dB) max.</th>
<th>Selectivity (dB) min.</th>
<th>Termination Impedance (ohms) max.</th>
<th>Spurious Response (dB) min.</th>
<th>Insertion Loss (dB) max.</th>
</tr>
</thead>
<tbody>
<tr>
<td>455±2 (±3KHz)</td>
<td>10</td>
<td>0</td>
<td>6 @ -10KHz</td>
<td>3K</td>
<td>10 (1-3MHz)</td>
<td>5</td>
</tr>
</tbody>
</table>

CFU 455

Figure 1.8: CFU455 Specifications

<table>
<thead>
<tr>
<th>Center Frequency (KHz)</th>
<th>6dB Bandwidth (KHz) min.</th>
<th>40dB Bandwidth (KHz) min.</th>
<th>Spurious Response (dB) min.</th>
<th>Insertion Loss (dB) max.</th>
<th>Input/Output Impedance (ohms)</th>
</tr>
</thead>
<tbody>
<tr>
<td>455±1.5</td>
<td>±10</td>
<td>±20</td>
<td>27</td>
<td>4</td>
<td>1500</td>
</tr>
</tbody>
</table>
1.4 High Gain 455 KHz I.F. Filter

In a superheterodyne receiver the I.F. filter is generally followed by amplification stages. The purpose of the I.F. amplifier is to bring the radio frequency (RF) signal that has been translated to the I.F. frequency up to an amplitude level appropriate for demodulation. The gain of the I.F. amplifier must also be adjustable so that varying signal strengths may be accommodated. Gain regulation is normally accomplished by using an automatic gain control (AGC) feedback loop. The high gain I.F. amplifier contained in the system box does not have this feature since the feedback path generally must come from the demodulator circuitry. Instead the system box circuit has a manual gain control (MGC) provided by a front panel potentiometer.

The schematic of the high gain 455KHz I.F. amplifier is shown in Figure 1.9. Note that the input is connected to the amplifier by first passing through a 455KHz ceramic filter of the type that was discussed earlier. The amplifier output is coupled to the output jack through a tuned transformer. The high gain I.F. amplifier will be used in the superheterodyne receiver experiment to allow reception of local AM broadcast stations.

![Figure 1.9: High Gain 455 KHz I.F. Amplifier](image-url)
1.5 Second-Order Active Lowpass Filter

In communication systems design, lowpass filters of one form or another are always required. The lowpass filter contained in the system box is an equal-component-value Sallen-Key circuit. The filter transfer function is given by

\[ H(s) = \frac{K\omega_o^2}{s^2 + \frac{\omega_o}{Q}s + \omega_o^2} \]  

(1.3)

where

\[ \omega_o = \frac{1}{RC} \]  

(1.4)

and

\[ Q = \frac{1}{3 - K} \]  

(1.5)

The quantity \( K \) is the noninverting amplifier gain which is given by

\[ K = 1 + \frac{R_a}{R_b} \]  

(1.6)

The second-order filter is implemented using a single op-amp as shown in Figure 1.10. The 39K and 33K resistors determine the filter gain and ripple characteristics. The series resistors each consisting of a fixed 1K resistor in series with a variable 100K resistor, are used to “tune” the filter cutoff frequency. A 10:1 resistance change provides a 10:1 frequency change.

\[ ^2 \text{F. W. Stephenson, RC Active Filter Design Handbook, Wiley, New York, 1985.} \]
1.6 Second-Order Active Highpass Filter

The highpass filter contained in the system box is the highpass transformed equivalent of the lowpass filter discussed above. The transfer function of the filter is given by

\[ H(s) = \frac{Ks^2}{s^2 + \frac{\omega_0}{Q}s + \omega_0^2} \]  

(1.7)

where \( \omega_0 \), \( Q \), and \( K \) are as defined for the lowpass transfer function. The schematic of this filter is shown in Figure 1.11. In the FM demodulation experiment the highpass filter will be used to implement a simple slope detection system.

1.7 Summing Amplifier/Envelope Detector

The variable gain summing amplifier with envelope detector output is just a single op-amp with a small signal diode in series with the output. The circuit schematic is shown in Figure 1.12. The transfer function characteristics of the amplifier are determined by connecting external components across the indicated terminals. The envelope detector characteristics are set by placing a parallel RC load across the detector output terminals.

In the communication systems experiments this circuit is used primarily for envelope detection of AM type signals. This includes demodulation techniques using carrier reinsertion (uses sum-
ming junction) as well as slope detection of FM (here the op-amp may be configured as a highpass filter). By shorting the diode the amplifier portion can be used as a loop filter in a phase-lock loop.

### 1.8 Audio Amplifier with Speaker Output

This portion of the system box contains an audio amplifier with input volume control and output connected to the speaker mounted on the front panel. The amplifier speaker combination is used for listening to various demodulated waveforms.
Chapter 2

Laboratory Experiments

This chapter of the lab manual contains the laboratory experiments that will be performed during the semester. The content of the experiments is designed to follow the lecture topics of ECE-4625 in such a way that concurrent enrollment should provide sufficient information to enable the student to understand the theoretical concepts of each lab. It is assumed however that the student has prior electronic circuit design laboratory experience such as ECE-3230 and/or ECE-3240.

At the present time the communications laboratory consists of six experiments:

- Linear Systems Characteristics
- Spectrum Analysis
- DSB and AM Modulation and Demodulation
- AM Superheterodyne Receivers
- Frequency Modulation and Demodulation
- Second Order Phase-Lock Loops

In the future more experiments are planned, and they will be added to this manual when appropriate. The experiments are contained in Sections 2.1 - 2.6.

2.1 Linear System Characteristics

The first part of this experiment will serve as an introduction to the use of the spectrum analyzer in making absolute amplitude measurements. The signals considered will be common periodic signals which are produced by a function generator (i.e. sine, square, and triangle waveforms). Next swept frequency measurements using a spectrum analyzer with a tracking generator will be investigated. The amplitude response of an active notch will be characterized. In the third portion of this experiment both the amplitude and phase response of linear systems will be considered. The use of a gain-phase meter will be introduced.
2.1. LINEAR SYSTEM CHARACTERISTICS

2.1.1 Basic Spectrum Analyzer Measurements

The spectrum analyzer is a device which displays an amplitude versus frequency plot of the frequency components or spectrum of any input signal. Located on roll-about carts, the lab has three Agilent 4395A spectrum/network analyzers. The analyzers cover the frequency range from 5Hz to 500MHz.

The operation of the 4395A is not as difficult as it might appear from just looking at the front panel controls. You will need to spend some time becoming familiar with both the the Network (quick start guide pp. 3-1 to 3-14) and Spectrum (quick start guide pp. 3-15 to 3-30) operating modes. Read through the Quick Start instructions carefully before proceeding any further with this experiment. For the first part of this experiment you will be using the analyzer in the Spectrum mode. When making measurements with this instrument the first step is to configure the operating state to the proper measurement mode, activate appropriate source and receiver ports, select frequency sweep parameters, and display formats. If you encounter difficulty or become frustrated trying to figure out the front panel controls, please ask the lab instructor for assistance.

The detailed front panel operation of both of these analyzers is described in the respective operators manual. A PDF version of the 4395A operators manual can be found on the course Web site (http://www.eas.uccs.edu/wickert/ece4670). Printed copies of the manuals can also be found on the respective carts. Ask your lab instructor to explain anything you don’t understand. The spectrum/network analyzer is a sophisticated instrument and can be easily damaged. DO NOT EXCEED THE MAXIMUM ALLOWABLE INPUT SIGNAL LEVELS (orange lettering below the input terminals). This instrument is very expensive, costing in excess of $28,000 new.

An overview of the spectrum analyzer measurement capabilities and input signal level maximums can be found in Table 2.1. An overview of the network analyzer measurement capabilities and input signal level maximums can be found in Table 2.2.

Table 2.1: Agilent 4395A spectrum analyzer mode general specifications and corresponding input maximums.

<table>
<thead>
<tr>
<th>Spectrum Analyzer Specifications</th>
<th>Attenuator Setting (dB)</th>
<th>Full Scale (max) Input (dBm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency Range 10 Hz to 500 MHz</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Noise Sidebands &lt; -104 dBc/Hz</td>
<td>0</td>
<td>-20</td>
</tr>
<tr>
<td>typical at 10 kHz offset</td>
<td>10</td>
<td>-10</td>
</tr>
<tr>
<td>Resolution Bandwidth 1 Hz to 1 MHz in 1-3-10 steps</td>
<td>20</td>
<td>0</td>
</tr>
<tr>
<td>Dynamic Range &gt; 100 dB third-order free dynamic range</td>
<td>30</td>
<td>+10</td>
</tr>
<tr>
<td>Level Accuracy ±0.8 dB @ 50 MHz -145 dBm/Hz</td>
<td>40</td>
<td>+20</td>
</tr>
<tr>
<td>Sensitivity -145 dBm/Hz @ freq. = 10 MHz</td>
<td>50</td>
<td>+30</td>
</tr>
</tbody>
</table>

The spectrum/network analyzer will first be used in the spectrum analyzer mode. Use of the network analyzer mode will be explored in the following section of the lab. In general spectrum
Table 2.2: Agilent 4395A network analyzer mode general specifications and corresponding input maximums.

<table>
<thead>
<tr>
<th>Network Analyzer Specifications</th>
<th>Attenuator Setting (dB)</th>
<th>Full Scale (max) Input (dBm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency Range</td>
<td>10 Hz to 500 MHz</td>
<td></td>
</tr>
<tr>
<td>Frequency Resolution</td>
<td>1 mHz</td>
<td></td>
</tr>
<tr>
<td>Output Power Range</td>
<td>-50 to +15 dBm</td>
<td>0</td>
</tr>
<tr>
<td>Dynamic Range</td>
<td>115 dB @ 10 Hz IFBW</td>
<td>10</td>
</tr>
<tr>
<td>Dynamic Accuracy</td>
<td>±0.05 dB/0.3 deg.</td>
<td>20</td>
</tr>
<tr>
<td>Calibration</td>
<td>Full two-port</td>
<td>30</td>
</tr>
<tr>
<td></td>
<td></td>
<td>40</td>
</tr>
<tr>
<td></td>
<td></td>
<td>50</td>
</tr>
</tbody>
</table>

Figure 2.1: Agilent 4395A Spectrum measurement test setup

and network analysis will be carried out as needed, throughout the entire series of lab experiments.

**Laboratory Exercises**

1. Using an Agilent 33120A or Agilent 33250A function generator connect a 1v p-p sine wave at 100KHz to the spectrum analyzer input as shown in Figure 2.1. The 1v p-p amplitude can be set using an oscilloscope provided the scope is set for 50Ω input impedance. The generator frequency can be set using an HP 5384A frequency counter.

   The measurements you will be taking are with respect to a 50 ohm impedance system. In a linear systems theory course a 1 ohm impedance environment is often assumed. This is convenient for mathematical modeling purposes, but in practice 50 ohms (75 ohms for cable TV) is the way radio frequency (RF) measurements are actually taken. The most common display mode for a spectrum analyzer is power in dB relative to one milliwatt. The units are denoted dBm. To find the power in dBm delivered to a 50 ohm load, from a single sinusoid
waveform, we might start with a scope waveform reading of volts peak-to-peak or peak.
From basic circuit theory, the power delivered to a load \( R \) is

\[
P_l = \frac{1}{2} \cdot \frac{v_{\text{peak}}^2}{R} = \frac{1}{2} \cdot \frac{(v_{\text{pp}}/2)^2}{R} = \frac{v_{\text{rms}}^2}{R} \text{ watts.} \quad (2.1)
\]

Recall that for a sinusoid, the rms value is \( v_{\text{peak}}/\sqrt{2} \), and clearly the peak value is one half the peak-to-peak value. In dBm the power is

\[
(P_l)_{\text{dBm}} = 10 \log_{10} \left[ \frac{P_l \text{ (watts)} \cdot 1000 \text{ mW/W}}{1 \text{ mW}} \right] = 10 \log_{10} \left[ P_R \text{ (watts)} \right] + 30 \quad (2.2)
\]

Suppose we have a 1v rms sinusoid in a 50 ohm measurement environment. The power level is

\[
(P_l)_{\text{dBm}} = 10 \log_{10} \left( \frac{1^2}{50} \right) + 30 = 13.01 \text{ dBm} \quad (2.3)
\]

Measure the input signal amplitude directly from the spectrum analyzer cursor numeric display using the following 4395A display modes:

(a) dBm (default)
(b) dBV
(c) VOLT

2. By observing an oscilloscope display, adjust the output of a function generator to obtain a 1v p-p square wave at 100KHz. Connect this signal to the spectrum analyzer. Note that you can “sharpen up” the frequency spike for each of the harmonics if you reduce the resolution bandwidth setting. On the Agilent 4395A analyzer this requires using appropriate front-panel key strokes. The quick start explains how to alter the resolution bandwidth (p. 3-23).
Ask your lab instructor for assistance if you are having problems.

(a) Identify and read the amplitudes of the various harmonics contained in the input signal. Compare your measured values with theoretical calculations obtained from Fourier analysis.
(b) Investigate the harmonic content of a triangle waveform, again compare the experimental results with theoretical calculations.

2.1.2 Vector Network Analyzer Measurements

In this portion of the experiment the 4395A will be used to characterize several linear systems. In particular linear two-port circuits will be analyzed in terms of the transmission parameters amplitude ratio (gain magnitude), phase, and group delay.

When making network analyzer measurements with this instrument the first step is to configure the operating state to the proper measurement mode, activate appropriate source and receiver ports, select frequency sweep parameters, and display formats. See the quick start guide pp. 3-1 to 3-14. If you encounter difficulty or become frustrated trying to figure out the front panel controls, please ask the lab instructor for assistance.
2.1. LINEAR SYSTEM CHARACTERISTICS

The front panel signal ports of the 4395A are designed for 50 ohm impedance level measurements (source and load resistances). A typical radio-frequency (RF) transmission measurement test setup is shown in Figure 2.2.

The 50 ohm power splitter provides equal amplitude signals with a 50 ohm source resistance to each of the splitter outputs. Receiver port $R$ is used to measure the signal at the input of the circuit under test while Receiver port $B$ measures the circuit output. With the measurement format set to Network: $B/R$ the complex ratio of these signal quantities as a function of sweep frequency is then formed by the analyzer. The analyzer screen can be configured to display two traces versus frequency, on a single plot or on a split plot (see the 4395A operators manual p. 6-3). For transmission measurements the typical display format would be the magnitude of the ratio in dB and the phase or group delay, which on the 4395A are display formats LOG MAG, PHASE, DELAY and $T/R(dB)$-τ).

For audio frequency circuits, which is the subject of this portion of the lab, we need a higher impedance level for making measurements. The setup shown in Figure 2.1b, which uses impedance transforming adapters or active probes on the receiver inputs, is more appropriate for audio measurements. The Agilent 41802A input adapter allows an ordinary 1MΩ scope probe to be used as an input device via an active network impedance transformation. The active probes HP 1124A transform the 50 ohm input impedance level to 10 MΩ shunted by 10 picofarads. The analyzer usable measurement bandwidth is reduced to about 100 MHz with both the 41802A and 1124A. The 1124A active probes are being replaced by the 41802A adapters. The specifications for the 41802A adapters is given in Table 2.3.
2.1. **LINEAR SYSTEM CHARACTERISTICS**

Table 2.3: Agilent 41802A input adapter specifications.

<table>
<thead>
<tr>
<th>41802A 1MOhm Input Specifications</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Frequency Range</strong></td>
</tr>
<tr>
<td><strong>Input Resistance</strong></td>
</tr>
<tr>
<td><strong>Input Capacitance</strong></td>
</tr>
<tr>
<td><strong>Accuracy</strong></td>
</tr>
<tr>
<td><strong>Gain Flatness</strong></td>
</tr>
<tr>
<td></td>
</tr>
<tr>
<td><strong>1 dB Gain Compression</strong></td>
</tr>
<tr>
<td><strong>Maximum Level</strong></td>
</tr>
</tbody>
</table>

Figure 2.3: RC Lowpass Filter

**Linear Circuits**

You will investigate two linear circuits using the vector network analyzer. The first is the familiar RC lowpass filter as shown in Figure 2.3. This circuit has no doubt received considerable attention in past lecture courses as well as electronic circuit laboratories. It is hoped that testing this circuit while you are becoming familiar with the network analyzer will make the interpretation of the experimental measurements easier.

For the second circuit you have your choice of analyzing an active bandpass filter (BPF) or an active bandstop filter (BSF). Both bandpass and bandstop filters are important in communication systems. The exact implementation technique of course depends on the particular application. Here we are assuming an audio frequency type application.

The design formulas for second order Sallen-Key BPF and BSF filters are provided below\(^1\). The circuit diagram of an equal component BPF is shown in Figure 2.4. The filter transfer function

2.1. LINEAR SYSTEM CHARACTERISTICS

Figure 2.4: Second Order Sallen-Key BPF

is given by

\[ H(s) = \frac{K\omega_0 s}{s^2 + \frac{\omega_0}{Q} s + \omega_o^2} \]  

(2.4)

where \( \omega_o \) is the filter center frequency, \( Q \) is the resonator quality factor related to the 3 dB bandwidth, and \( K \) is the amplifier voltage gain. The design equations are

\[ \omega_o = \frac{\sqrt{2}}{RC} \]  

(2.5)

\[ Q = \frac{\sqrt{2}}{4 - K} = \frac{f_o}{BW} \]  

(2.6)

\[ K = 1 + \frac{R_a}{R_b} \]  

(2.7)

Note that the filter gain at center frequency is given by

\[ H(j\omega_o) = \frac{K}{4 - K} \]  

(2.8)

A general BPF frequency response showing the relation between \( Q \) and BW is shown in Figure 2.5a. The schematic of the Sallen-Key BSF is shown in Figure 2.6. The filter transfer function is given by

\[ H(s) = \frac{K(s^2 + \omega_o^2)}{s^2 + \frac{\omega_0}{Q} s + \omega_o^2} \]  

(2.9)

The parameters \( \omega_o \), \( Q \), and \( K \) have the same meaning as in the BPF case except the design equations for \( = \omega_o \) and \( Q \) are now

\[ \omega_o = \frac{1}{RC} \]  

(2.10)

\[ Q = \frac{1}{4 - 2K} \]  

(2.11)
2.1. LINEAR SYSTEM CHARACTERISTICS

Figure 2.5: BPF and BSF Frequency Response and Bandwidth Definitions

Figure 2.6: Second Order Sallen-Key BSF
The general BSF frequency response and bandwidth definitions are shown in Figure 2.5b.

Laboratory Exercises

1. Lowpass Filter

   (a) Referring to the lowpass filter of Figure 2.3 using \( C = 1 \) nF or another suitable value, determine \( R \) such that the cutoff frequency is near 10 kHz.

   (b) Using the 4395A configured for vector network transmission measurements, frequency sweep the lowpass filter from 10 Hz to 100 KHz to obtain gain in dB, phase in degrees, and group delay in microseconds versus frequency. Plot and label your experimental data using the screen capture capability of the 4395A. You may wish to investigate the marker/line cursor menus for analyzing your experimental results.

   (c) Derive the corresponding theoretical expressions for gain, phase and group delay and then compare your experimental data with theoretical predictions. Consider also an Agilent Advanced Design System (ADS) simulation using the AC analysis controller. Your lab instructor will give you an introduction to ADS.

   (d) Attempt to justify any significant deviations from your theoretical predictions. For example the lumped element RC filter model could be enhanced by including the parasitic capacitance between terminal strips of the breadboard.

2. Bandpass or Bandstop Filter

   (a) Design and build either a BPF or a BSF as shown in Figure 2.4 and Figure 2.6 respectively. Use an LF-353 or similar (bi-fet) op-amp. Design the filter to have a center frequency of about 10 kHz and \( 2 \leq Q \leq 5 \). As a starting point you may wish to choose \( C = 1 \) nf. To allow a variable \( Q \) you may also wish to replace the gain set resistors, \( R_a \) and \( R_b \), with a potentiometer (10 k ohm works fine).

   (b) Using the 4395A configured for network transmission measurements, frequency sweep the active filter circuit over a range of frequencies centered on \( \omega_o \). Measure filter performance data corresponding to the design parameters. The gain in dB is of primary importance here. Again use the screen capture capabilities of the 4395A for obtaining a hard copy of your data.

   (c) Compare the measured amplitude response with the design parameters and comment on deviations from theory.

   (d) To obtain a better understanding of the theoretical circuit performance simulate the circuit using ADS or Spice. The lab PC’s have ADS installed. A tutorial notes and screencasts on using ADS can be found on the ECE 5250/4250 Web Site (soon on the course Web site).

   (e) Why is it impractical to obtain high \( Q \)’s from these circuits?
2.1.3 System Box Filter Characterization

In this final portion of the experiment, the active lowpass and highpass filters that are part of the system box will be characterized. This information will be helpful in future experiments. The instrumentation used for obtaining performance results can be either the 4395A Network Analyzer. The magnitude response is of primary importance at this time.

Laboratory Exercises

Using the 4395A:

1. Determine the range of 3 dB cutoff frequencies for the active lowpass filter.
2. Verify that the slope of the LPF transition band corresponds to a second order response.
3. Determine the range of 3dB cutoff frequencies for the highpass filter.
4. Verify that the slope of the HPF transition band corresponds to a second order response.
2.2 Spectrum Analysis

In the study of communication circuits and systems, frequency domain analysis techniques play a key role. The communication systems engineer deals with signal bandwidths and signal locations in the frequency domain. The tools for this analysis are Fourier series and Fourier transforms. The types of signals encountered can range from simple deterministic signals to very complex random signals. In this experiment only deterministic signals which are also periodic will be considered.

2.2.1 Background

Power Spectral Density of Periodic Signals

The frequency domain representation of periodic signals can be obtained using the complex exponential Fourier series. If \( x(t) \) is periodic with period \( T_0 \), that is \( x(t) = x(t + kT_0) \) for any integer \( k \), then

\[
x(t) = \sum_{n=-\infty}^{\infty} X_n e^{j2\pi n f_0 t}
\]

where

\[
X_n = \frac{1}{T_0} \int_{T_0} x(t) e^{-j2\pi n f_0 t} dt
\]

and \( f_0 = 1/T_0 \).

An alternate approach is to take the Fourier transform of \( x(t) \) using the concept of Fourier Transforms in the limit. To develop this approach begin by letting \( x(t) \) be periodic, then write

\[
x(t) = \sum_{m=-\infty}^{\infty} p(t - m T_0)
\]

where

\[
p(t) = \begin{cases} x(t) & |t| \leq T_0/2 \\ 0 & \text{otherwise} \end{cases}
\]

To find \( X(f) \) the following transform pair can then be used

\[
\sum_{m=-\infty}^{\infty} p(t - m T_0) \leftrightarrow \sum_{n=-\infty}^{\infty} f_0 P(n f_0) \delta(f - n f_0)
\]

where

\[
P(f) = \mathcal{F}\{p(t)\}
\]

A useful frequency domain representation of a power signal \( x(t) \) is the power spectral density \( S_x(f) \). \( S_x(f) \) gives the distribution of power in \( x(t) \) versus frequency. Since \( S_x(f) \) is a density it has units of W/Hz. The power spectral density is defined as

\[
S_x(f) = \mathcal{F}\{R_x(\tau)\} = \int_{-\infty}^{\infty} R_x(\tau) e^{-j2\pi f \tau} d\tau
\]
where $R_x(\tau)$ is the autocorrelation function of the signal $x(t)$. The autocorrelation function is computed as a time average for deterministic signals, and a statistical average for random signals. The time average definition of $R_x(\tau)$ is

$$R_x(\tau) = \lim_{T \to \infty} \frac{1}{2T} \int_{-T}^{T} x(t)x(t + \tau)dt$$

$$= \frac{1}{T_0} \int_{T_0} x(t)x(t + \tau)dt\text{ for } x(t)\text{ periodic} \quad (2.19)$$

If $x(t)$ is periodic it then follows that $R_x(\tau)$ is also periodic which implies that $S_x(f)$ will contain impulses. For $x(t)$ real and periodic it is a simple matter to show that $R_x(\tau)$ and $S_x(f)$ can be written in terms of the Fourier coefficients corresponding to $x(t)$. The result is

$$R_x(\tau) = \sum_{n=-\infty}^{\infty} |X_n|^2 e^{j2\pi nf_0\tau} \quad (2.20)$$

and

$$S_x(f) = \sum_{n=-\infty}^{\infty} |X_n|^2 \delta(f - nf_0) \quad (2.21)$$

**Pseudo-Noise Sequence Generators**

In the testing and evaluation of digital communication systems a source of *random like* binary data is required. A maximal-length sequence generator or pseudo-noise (PN) code is often used for this purpose. A generator of this type produces a sequences of binary digits that is periodic. For a listing of the properties of maximal-length sequences see Ziemer and Peterson\(^2\). Typical uses of PN sequences are:

- **Encipherment**: A message written in binary digits may be added modulo-2 (exclusive-OR) to a PN sequence acting as a key. The decipherment process consists of adding the same PN sequence, synchronized with the first, modulo-2 to the encoded message. Such a technique is a form of scrambling.

- **Randomizer**: A PN sequence can also be used for breaking up long sequences of 1’s or 0’s which may bias a communications channel in such a way that performance of the communications system is degraded. The PN sequence is referred to as a randomizer. Note that the scrambler mentioned above and the randomizer perform the same function, but for different purposes.

- **Testing the performance of a data communication system.**

- **Code generator.**

- **Prescribed Sequence Generator** such as needed for word or frame synchronization.

---

Practical implementation of a PN code generator can be accomplished using an N-stage shift register with appropriate feedback connections. An $N$-stage shift register is a device consisting of $N$ consecutive binary storage positions, typically D-type flip-flops, which shift their contents to the next element down the line each clock cycle. Without feedback the shift register would be empty after $N$ clock cycles. A general feedback configuration is shown in Figure 2.7 with the input to the first stage being a logical function of the outputs of all $N$ stages. The resulting output sequence will always be periodic with some period $M$. The maximum sequence period is $2^N - 1$. The feedback arrangement which achieves the maximum sequence period results in a maximum length sequence or $m$-sequence. The block diagram and output sequence for a three stage $m$-sequence generator are shown in Figure 2.8. The logic levels here are assumed to be bipolar ($\pm A$). In practice the actual logic levels will correspond to the device family levels. Possible exclusive-OR feedback connections for $m$-sequences with $N = 1$ through $N = 15$ are given in Table 2.4.

The power spectrum $S_x(f)$ of the $m$-sequence generator output can be found from the auto-correlation function $R_x(\tau)$. By definition we can write

$$ R_x(\tau) = \frac{1}{MT} \int_{-MT/2}^{MT/2} x(t)x(t + \tau) dt \quad (2.22) $$

where $T$ is the clock period. For long sequences the above integral is difficult to evaluate.

An example of how to compute $R_x(\tau)$ for $M = 7$ is given below. For this example a 0 is used in place of a $-1$ for the bit ‘0’, but multiplication is defined as $0 \times 1 = 1 \times 0 = -1$ and $0 \times 0 = 1 \times 1 = 1$ which is a bipolar AND operation. We will find the autocorrelation function at discrete times $\tau = kT$ using the normalized correlation

$$ R_x(kT) = \frac{1}{M} \sum_{i=1}^{M} \rho_i \quad (2.23) $$

where $\rho_i$ is the product of a bit in the sequence with a corresponding bit in a shifted version of the sequence. The calculation of $R_x(kT)$ for $k = 0, 1, and 2$ is shown in Figure 2.9. The continuous autocorrelation function follows by connecting the discrete time points together as shown in Figure 2.10.
2.2. SPECTRUM ANALYSIS

Clock
Period = T

\[
M = 2^3 - 1 = 7
\]

Figure 2.8: Three Stage m-Sequence PN Generator

Table 2.4: m-Sequence Generator Feedback Connections

<table>
<thead>
<tr>
<th>Stages</th>
<th>Connections</th>
<th>Stages</th>
<th>Connections</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>(Q_1)</td>
<td>9</td>
<td>(Q_5 \oplus Q_9)</td>
</tr>
<tr>
<td>2</td>
<td>(Q_1 \oplus Q_2)</td>
<td>10</td>
<td>(Q_7 \oplus Q_{10})</td>
</tr>
<tr>
<td>3</td>
<td>(Q_2 \oplus Q_3)</td>
<td>11</td>
<td>(Q_9 \oplus Q_{11})</td>
</tr>
<tr>
<td>4</td>
<td>(Q_3 \oplus Q_4)</td>
<td>12</td>
<td>(Q_2 \oplus Q_{10} \oplus Q_{11} \oplus Q_{12})</td>
</tr>
<tr>
<td>5</td>
<td>(Q_3 \oplus Q_5)</td>
<td>13</td>
<td>(Q_1 \oplus Q_{11} \oplus Q_{12} \oplus Q_{13})</td>
</tr>
<tr>
<td>6</td>
<td>(Q_5 \oplus Q_6)</td>
<td>14</td>
<td>(Q_2 \oplus Q_{12} \oplus Q_{13} \oplus Q_{14})</td>
</tr>
<tr>
<td>7</td>
<td>(Q_6 \oplus Q_7)</td>
<td>15</td>
<td>(Q_{14} \oplus Q_{15})</td>
</tr>
<tr>
<td>8</td>
<td>(Q_2 \oplus Q_3 \oplus Q_4 \oplus Q_8)</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
### 2.2. SPECTRUM ANALYSIS

<table>
<thead>
<tr>
<th>Sequence</th>
<th>← one period →</th>
</tr>
</thead>
<tbody>
<tr>
<td>0 1 1 1</td>
<td>0 0 1 0 1 1 1</td>
</tr>
<tr>
<td>Correlation</td>
<td>1 1 1 1 1 1 1</td>
</tr>
<tr>
<td>$R_x(0) = 1$</td>
<td></td>
</tr>
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</table>

<table>
<thead>
<tr>
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<th>← one period →</th>
</tr>
</thead>
<tbody>
<tr>
<td>0 1 1 1</td>
<td>0 0 1 0 1 1 1</td>
</tr>
<tr>
<td>Seq. Shifted Once 1 0 1 1</td>
<td>1 0 0 1 0 1 1</td>
</tr>
<tr>
<td>Correlation</td>
<td>-1 -1 -1 -1 -1 -1 -1</td>
</tr>
<tr>
<td>$R_x(0) = -1/7$</td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Sequence</th>
<th>← one period →</th>
</tr>
</thead>
<tbody>
<tr>
<td>0 1 1 1</td>
<td>0 0 1 0 1 1 1</td>
</tr>
<tr>
<td>Seq. Shifted Twice 1 1 0 1</td>
<td>1 1 0 0 1 0 1</td>
</tr>
<tr>
<td>Correlation</td>
<td>-1 -1 -1 -1 -1 -1 -1</td>
</tr>
<tr>
<td>$R_x(0) = -1/7$</td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Sequence</th>
<th>← one period →</th>
</tr>
</thead>
<tbody>
<tr>
<td>0 1 1 1</td>
<td>0 0 1 0 1 1 1</td>
</tr>
<tr>
<td>Seq. Shifted Three 1 1 0 0</td>
<td>1 0 1 1 1 0 0</td>
</tr>
<tr>
<td>Correlation</td>
<td>-1 -1 -1 -1 -1 -1 -1</td>
</tr>
<tr>
<td>$R_x(0) = -1/7$</td>
<td></td>
</tr>
</tbody>
</table>

Figure 2.9: M=7 Sequence Correlation Function Calculation

![Figure 2.10: M=7 Autocorrelation Function](image)

Figure 2.10: M=7 Autocorrelation Function


2.2. SPECTRUM ANALYSIS

Figure 2.11: Periodic Autocorrelation Function

2.2.2 Preliminary Analysis

1. Find the power spectral density of a periodic pulse train signal

\[ x(t) = \sum_{m=-\infty}^{\infty} A \Pi \left( \frac{t - mT_s}{\tau} \right) \quad 0 < \tau < T_s \quad (2.24) \]

2. Find the power spectral density of a periodic cosinusoidal pulse train

\[ x(t) = \sum_{m=-\infty}^{\infty} A \Pi \left( \frac{t - mT_s}{\tau} \right) \cos 2\pi f_o t \quad 0 < \tau < T_s \quad (2.25) \]

3. Find the power spectral density corresponding to the periodic autocorrelation function shown in Figure 2.11.

2.2.3 Periodic Pulse Train

Laboratory Exercises

Using an Agilent 33250 generator record time domain and frequency domain data using the oscilloscope and spectrum analyzer respectively, for several different duty cycles \( \tau / T_s \). As a matter of convenience select \( T_s \) as an integer number of graticule lines on the oscilloscope screen. Pressing the Pulse button on the front panel of the 33250A puts the generator in pulse train mode. A good starting point is a low amplitude of 0 v and a high amplitude of 200 mv (200 mv p-p). The frequency \( F_s \) can be 10 kHz, making \( T_s = 1 / f_s = 100 \mu s \). Then choose pulse width \( \tau = 10 \mu s \).

The frequency domain data should include:

1. The spacing between spectral lines.
2. The location of zeros in the spectrum envelope.
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3. The number of spectral lines between zeros.

4. Any additional data that seems appropriate for comparing the experimental power spectral density with theoretical calculations. The VisSim/Comm simulation file lab2a.vsm, with screen shot shown in Figure 2.12 and available on the course web site, can also be used to help correlate measured results with theoretical results.

5. The pulse rise and fall time can be altered on the 33250A by changing the Edge Time, which has a minimum/default value of 5 ns. Try increasing this up to 25% of the pulse width, you should now have a trapezoidal shaped pulse. Comment on the change in spectral content at high frequencies as the rise and fall times increase. Is this the expected result? The VisSim/Comm simulation file lab2a_finite.vsm and shown in Figure 2.13, is also available for comparison with measured results.

2.2.4 Periodic Cosinusoidal Pulse Train

A periodic cosinusoidal pulse train can easily be generated using the Agilent 33250A in Burst Mode.
2.2. SPECTRUM ANALYSIS

Laboratory Exercises

1. The waveform should be of the form shown in Figure 2.14. Start with a 200 mv p-p amplitude and \( f_o = 1/T_o = 500 \) kHz. The pulse repetition rate should be \( f_s = 1/T_s = 10 \) kHz, or for 33250A you set burst period = 100 \( \mu s \). On the 33250A the value for pulse width \( \tau \) is set in terms of the number of cycles per burst, that is \( \tau = M T_o \) provided \( M T_o < T_s \). Start with \#cycles = 10 which makes \( \tau = 20 \) \( \mu s \).

2. Take frequency domain data from the spectrum analyzer similar to that of part 1, above, so that the experimental power spectral density can be compared with theoretical calculations and simulation results. The VisSim/Comm simulation file lab2b.vsm, shown in Figure 2.15 can be used for comparison purposes to your measurements and your theoretical expectations.

3. Observe the spectra for different values of \( \tau/T_s \) with \( f_o \) fixed. Comment on your observations.

4. Now vary \( f_o \) about 500 kHz and observe the spectra. Comment on your observations.
2.2. SPECTRUM ANALYSIS

Note on the 33250A: $\tau = MT_o$, $T_s > MT_o$

Figure 2.14: Coherent Cosinusoidal Pulse Train Waveform

Figure 2.15: VisSim/Comm simulation file for cosinusoidal pulse train simulation
2.2.5 Pseudo-Noise (PN) Sequence Generators

In this portion of the experiment you will verify some of the properties of PN sequence generators, specifically m-sequences. In the lab you will find an assembled PN generator which is constructed using TTL logic family parts. The circuit consists of a switch selectable 5/10 stage feedback shift register with a built-in automatic start feature in the event that the shift register powers up in the all zeros state. The schematic drawing of the generator is shown in Figure 2.16. The clock input is standard TTL as are all of the outputs. The sync output should be connected to the scope external trigger input.

Laboratory Exercises

1. With the PN generator set to the 5 stage position verify that the sequence length is as predicted.

2. Using the spectrum analyzer verify that the general shape of the power spectral density you observe corresponds to that which you calculated earlier.

3. With the generator still in the 5 stage mode record the clock frequency, the spectral line spacing, and the spacing between the spectral envelope zeros. Use this data to make comparisons with theory. Note to observe spectral line spacing it may be necessary to manually reduce the spectrum analyzer resolution bandwidth (Agilent 4395A).

4. Compare your results with VisSim/Comm simulation results produced by 1ab2c.vsm and shown in Figure 2.17.

5. Repeat 1–4 above with the generator in the 10 stage position. Note: obtaining the proper scope triggering may be difficult. Both spectrum analyzer are capable of resolving closely spaced sinusoids, so with the longer code period, reducing the resolution bandwidth will be even more critical.
2.2. SPECTRUM ANALYSIS

Figure 2.16: 5/10 Bit m-Sequence Generator
2.2. SPECTRUM ANALYSIS

Figure 2.17: VisSim/Comm simulation file for PN generator simulation
2.3 DSB and AM Modulation and Demodulation

In this experiment linear analog modulation techniques will be investigated along with appropriate demodulation schemes. The first part of this experiment will allow the student to become familiar with the operation of the 1496 balanced modulator integrated circuit, which is part of the system box. A DSB transmitter is then constructed along with both coherent and non-coherent demodulators. Following this AM signal generation and envelope detection is investigated.

2.3.1 Coherent Demodulation of DSB

To construct a DSB modulator and demodulator a signal processing function which performs multiplication is needed. The Motorola MC 1496 balanced modulator chip serves this purpose in the communications experiments. The equivalent block diagram for this circuit as shown in Figure 2.18, has a signal or message input, \( V_s \), a carrier input, \( V_c \), and a balance control pot. The balance control performs a function that may be represented mathematically as adding a dc voltage to the message signal.

![Multiplier Block Diagram](image)

If the message signal \( V_s = m(t) \) is added to a dc voltage of value \( A \), then we may express the output as

\[
V_o = [A + V_s] V_c = [A + m(t)] V_c \\
= [A + m(t)](B \cos \omega_c t) \\
= AB \cos \omega_c t + m(t)B \cos \omega_c t
\]

In this expression, \( AB \cos \omega_c t \) represents the carrier frequency component of the frequency spectrum for the modulated waveform. Since \( A \) is a dc voltage which can be adjusted by the balance control, we see that if we let \( A = 0 \), the carrier frequency component is “balanced out”. This condition is described as suppressed carrier modulation or sometimes double side-band.

Multiplier Familiarization Exercises

Setup the system shown in the block diagram given in Figure 2.19. Use the Agilent 33250A and/or Agilent 33120A function generators for the sources. Later when you need a second carrier source for the receiver, you may want to use one of the older function generators, e.g., HP3312A or Krohn-Hite 1200A, for the message source.
1. Using an oscilloscope adjust the sine wave carrier amplitude to about 200mv peak-to-peak and adjust the message source to about 1v peak-to-peak, with the balance control near the center of its range. Set the scope input to AC coupling and synchronize to the message waveform.

2. To observe the limits on the modulator performance and to help you in setting up your experiments, do the following:

   (a) Adjust the balance control through its full range of travel. Notice that the waveform is clipped of flattened out at the top and bottom as you reach the saturation level of the modulator output amplifier.

   (b) With a triangle waveform for $m(t)$, adjust the balance control again. You can see the waveform distortion more easily in this case, since the point of the triangle is more distinctive than the top of a sinusoidal waveform.

   (c) Adjust the balance control until approximately one-half of the message waveform is flattened out. Observe now that by increasing the level of $V_c$, which adds more carrier, the distortion remains the same.

   (d) Return the carrier to 200mv peak-to-peak. Observe now that the distortion in the modulated waveform can be reduced or eliminated by changing the signal level, $V_s$.

   (e) In every case, for any given modulation index, a distortion in the envelope of the modulated waveform can only be eliminated by reducing the signal level, $V_s$, and not by changing carrier level. It is possible, however, to have a distortion in the carrier itself, which will produce harmonics in the frequency spectrum at multiples of the carrier frequency. To observe this effect, reduce the message to about zero amplitude and adjust the balance so no envelope distortion is evident. Now increase the carrier level, $V_c$, until a hardening or brightening of the top and bottom edges of the waveform is evident. By synchronizing the scope such that individual cycles of the carrier are seen, the distortion of the carrier sine wave will be observed as the carrier level is increased. Also observe the results on the spectrum analyzer. To connect the Agilent 4395A analyzer to the modulator output use a high impedance active probe. Obtain probe power from the jack on the front of the analyzer. The probe input impedance is $1 \, M\Omega$ compared
with the 50 Ω impedance of the straight analyzer input. This avoids circuit loading problems.

(f) Return the carrier to 200mv peak-to-peak and the modulating signal to 1v peak-to-peak or less so that distortion is not seen. Now note that the modulator output has a considerable dc level added to the signal. Observe this by switching the scope input from AC to DC, and also by measuring the modulated waveform with a dc voltmeter. Explain in your notes how the dc voltmeter gets an accurate measurement from the composite waveform.

This dc level **must not be ignored**. In connecting the output of the modulator into any system, the dc must be removed by level shifting, by capacitive coupling, or in some other manner if you are going to obtain proper results and prevent damage to equipment.

At this time you have seen all of the features and problems of the IC modulator (multiplier) circuit (MC 1496) used in the lab. You are now prepared to proceed with the communications experiments and will be expected to know how to use and interpret the output signals you see.

**Laboratory Exercises**

1. Adjust the 10 kHz message source to near the maximum possible without distortion. Connect the modulator output to the op-amp level shifter contained in the system box. Adjust the offset control until the op-amp output is centered about zero, as measured with a dc voltmeter and observed on a scope.

2. Reduce the modulator input, $V_s$, of the message source to zero by using the output attenuator on the function generator to make the signal amplitude very small. Now balance out (see page 35) the carrier by adjusting the multiplier DC offset control and observing the output on a scope. Bring the message amplitude back to its original setting.

3. Now connect the level shifter output to the $V_s$ input of a second multiplier and construct the circuit shown in Figure 2.20. This circuit is a simulation of a synchronous or coherent demodulation system. The circuit constructed during the multiplier familiarization exercises
represents the radio transmitter with output \( s(t) \) being the modulated radio frequency wave. Normally the signal is received at an antenna, amplified, and *mixed* (multiplied) with a local oscillator at the receiver in order to demodulate or recover the original signal, \( m(t) \).

**For Your Notes:** The signal \( s(t) \) can be expressed as

\[
s(t) = m(t) c(t) = m(t) \cos \omega_c t
\]  

(2.27)

Write an expression for the signal \( y(t) = s(t) c(t) \). Show, using trig identities, that \( y(t) \) is proportional to the sum of the original message and a modulated double frequency carrier. Sketch a cycle of the sine wave \( m(t) \) and the corresponding modulated carrier with the proper relative amplitudes and sketch the sum of the two waveforms. Sketch the frequency spectra of \( s(t) \), \( c(t) \), and \( y(t) \).

4. Adjust the amplitude of signals going to the multipliers, if necessary, to obtain a clean undistorted waveform for \( y(t) \).

5. Connect \( y(t) \) to the second order lowpass filter in the system box. Since the lowpass filter is direct coupled, a coupling capacitor or op-amp level shifter must be used to remove the dc offset present at the output of the second multiplier.

6. Over what range of frequencies should the filter cutoff frequency, \( f_{3dB} \), be chosen from?

7. Now observe on a scope and spectrum analyzer both the filter input and output. If the filter output looks like a *fuzzy* version of \( m(t) \), then you know that the filter bandwidth is too wide. For the system to be useful as a demodulator for more general signals, the frequency at which the *fuzz* disappears from the signal should be above the signals highest frequency or message bandwidth.

8. To gain additional understanding compare the waveforms and spectra obtained from the hardware measurements with the VisSim/Comm simulation shown below in Figure 2.21, and found in file lab3a.vsm.

9. In the steps above you have simulated a coherent or synchronous receiver for an AM-SC, or DSB signal. By way of proving to yourself that exact synchronism between the transmitter and receiver oscillators is required, substitute another function generator for the demodulator oscillator and see if you can tune it well enough to reliably recover the \( m(t) \) waveform. When you get close you will see a clean, undistorted version of \( m(t) \) that fades in and out rather rapidly. Note to be able to obtain reasonably stable synchronism, it is recommended that the transmitter receiver oscillators be chosen from the synthesized function generators, Agilent 33250A and Agilent 33120A.

10. **For Your Notes:** Express \( y(t) \) as the product of the signal \( s(t) \) and a local oscillator \( c'(t) \) that has the same frequency as the original carrier \( c(t) \), but a time varying phase relationship with respect to \( c(t) \). That is

\[
y(t) = s(t) c'(t) = s(t) \cos[\omega_c t + \phi(t)]
\]  

(2.28)
2.3. DSB AND AM MODULATION AND DEMODULATION

Double Sideband Mod/Demod

Transmit Multiplier

500 kHz Cosine (1 V)

Receiver Multiplier

500 kHz Cosine (0.1 V)

Local Reference

Demodulator

Lowpass Filter

Local Receiver Reference

lab3a.vsm

Time Domain Plots

Transmitted DSB

Demodulated DSB Before LPF

Demodulated DSB After LPF

Time Domain Plots

Frequency Domain Plots

Transmitted DSB

Demodulated DSB Before LPF

Demodulated DSB After LPF

Frequency Domain Plots

Figure 2.21: VisSim/Comm simulation file for DSB modulation and coherent demodulation
2.3. DSB AND AM MODULATION AND DEMODULATION

Expand this expression using trig identities and use the results to explain your observations in the non-synchronous demodulation of DSB. Use the VisSim/Comm simulation shown in Figure 2.22, file lab3b.vsm, to further explain your theoretical analysis.

11. Using an FM radio as a source, \( m(t) \), for voice/music signals, set up the multipliers and filters as shown in Figure 2.19. Connect the output of the lowpass filter to the system box audio amplifier so that first the single tone message can be heard through the speaker, then use the FM radio source. To obtain an audio output signal from the FM radio use the headphone jack, which also mutes the speaker. Observe the effects noted earlier in synchronous and non-synchronous demodulation of DSB.

### 2.3.2 Diode Envelope Detector for an AM Signal

When a modulated signal is formed by multiplying a strictly positive waveform with a carrier signal, the result is an amplitude modulated waveform whose envelope contains the transmitted information. That is,

\[
s(t) = [A + m(t)] B \cos \omega_c t \tag{2.29}
\]

is an AM waveform as described above if the dc level \( A \) is greater than the peaks of \( m(t) \), so that \( A + m(t) \) is never negative.

The ratio of the peak value of \( m(t) \) to the dc level \( A \) is known as the modulation index, \( a \), or

\[
a = \frac{|m(t)|_{\text{peak}}}{A} \tag{2.30}
\]
This number must be less than or equal to one to satisfy the requirement that \( A + m(t) \) be positive.

A synchronous demodulator can be used to recover the information, \( m(t) \), from a modulated waveform, but this requires a synchronized local oscillator and is an unnecessarily complicated detector for this type of signal. If the modulation index is less than one, a simple diode detector may be used as shown in Figure 2.23. In this circuit the capacitor charges up to the peak value of each cycle of the carrier waveform. Between cycles the capacitor discharges through the resistor with time constant depending upon \( R \) and any load on the detector.

**Laboratory Exercises**

1. The Summing Amplifier/Envelope Detector block contained in the system box can be used to implement the diode detector shown in Figure 2.23. The op-amp serves as a buffer and a gain block. Place input and feedback resistors of appropriate values around the op-amp, and a parallel RC network across the diode output. Set the op-amp gain magnitude between 10 and 100. The RC time constant can be set using resistor and capacitor decade boxes or fixed components.

2. To obtain an AM source use the DSB transmitter circuit from the first part of this experiment, except now the DC offset control will be used to control the modulation index. An alternate AM source is the Agilent 33250A (preferred) or Agilent 33120A function generator, with the internal modulation generator set for AM. In either case adjust an unmodulated 500 KHz carrier to about 1 v peak-to-peak. With a message frequency of approximately 1 kHz, generate an AM wave with a modulation index of 0.5. Connect the AM source to the input of the envelope detector. Observe the output on the scope synchronized to the message waveform. You may want to use an external sync signal for the scope. Ensure that the modulated waveform has a slightly negative or exactly zero dc offset (use the dc voltmeter). Alternatively, use a coupling capacitor to block any dc component.
To obtain $\alpha = 0.5$ note that the equation for the AM wave may also be written as

$$s(t) = A \left[1 + \frac{|m(t)|}{A}\right] \cos \omega_c t$$

where the peak of $|m(t)|$ divided by $A$ is the modulation index. If the modulation index is given as 0.5, this means that the unmodulated carrier is increased by 50% and decreased by 50% by the peaks of the modulating waveform. Set the unmodulated carrier to four divisions in height on the scope. Now adjust the modulation so that the carrier is increased and decreased by one division. This is 50% modulation. Basically, then, the percentage increase or decrease in the carrier from its unmodulated condition is the percentage of modulation.

3. Using various values of resistance and capacitance, observe the operation of the envelope detector. Use capacitors from about 500 pf to 0.01 $\mu$F. Calculate the RC time constant necessary to just follow the steepest slope of a 4$\mu$V peak-to-peak sinusoidal waveform at 1 kHz. Calculate $C$ if $R = 10k$. The optimum value should be $RC \approx 1/\omega_{\text{max}}$.

4. Try to pick a best value of $R$ and $C$ to track a 1 kHz modulating signal. Then reduce and increase the signal frequency and observe the effect. (Your final values should be in the neighborhood of 10k and 0.01$\mu$F. A general relationship is to choose $RC \approx 1/\omega_{\text{max}}$.)

5. Reduce and increase the modulation index. Can you explain why the waveform bottoms out at a high modulation index? Hint: Consider the diode forward voltage drop required for conduction. Is it possible for a circuit like this to ever recover a signal from a 100% modulated waveform? Explain (recall the super diode circuit that is discussed in electronic circuits).

6. Compare your hardware results with the VisSim/Comm simulation shown in Figure 2.24, file lab3c.vsm. Note that in this simulation file there are two representations of the diode envelope detector.

7. Connect the audio amplifier to the envelope detector output and listen to the recovered message signal using both the sinusoid and FM radio music source. Vary the modulation index and observe the effects.

### 2.3.3 RF Trapezoid and Modulation Index

For modulation signals which are symmetrical in the sense that they have equal magnitude positive and negative peaks, such as a sine wave, the modulation index can be determined rather easily using a scope. By sweeping an AM waveform across the scope CRT using the message signal itself, a trapezoid display is obtained.

**Laboratory Exercises**

1. Adjust an AM waveform to about 50% modulation with several cycles of the modulating signal displayed on the scope.
2.3. DSB AND AM MODULATION AND DEMODULATION

Figure 2.24: VisSim/Comm simulation file for AM modulation and envelope detection
2. Connect the modulating signal, \( m(t) \), to channel 1 of the scope and the modulated signal to channel 2. Next press the button labeled Main/Delayed and then the x-y soft key. Adjust the gain controls of both scope input channels until you see a full screen width trapezoid shaped waveform. Trigger on channel 2, adjusting the trigger level close to the waveform top.

3. Adjust the AM signal modulation index until the trapezoid comes to a point. Now switch back to the calibrated time base horizontal sweep. You should have a 100% modulated waveform. Observe the time function and the trapezoid for the overmodulated and several other cases, explain in your notes, with words and sketches, how the modulated carrier wave and the trapezoid pattern are related.

4. Switch back to the trapezoidal waveform and determine how to obtain the modulation index from this pattern. Show how the peaks and valleys of the modulated waveform are related to the ends of the RF trapezoid, and show how to obtain the modulation index, \( a \), algebraically by using the equations for the length of the ends of the trapezoid.

**Hint:** An incorrect formula that works for the limit cases of \( a = 1 \) and \( 0 \) may be avoided if you also check your formula for \( 0 < a < 1 \) as well, e.g., \( a = 0.5 \).
2.4 AM Superheterodyne Receivers

The block diagram of a basic superheterodyne receiver is illustrated in Figure 2.25. In a typical broadcast radio receiver, the input to the RF amplifier is obtained from the antenna tuning circuit as the station selector dial changes the capacitance and therefore the resonant frequency of the LC tank circuit. The local oscillator frequency is adjusted simultaneously by varying a tuning capacitor that is also connected to the station selector dial. The local oscillator frequency is designed to change so that it is always a fixed value above the selected station frequency. Thus, if the RF tuner is centered at a frequency \( f_c \), then

\[
f_{LO} = f_c + f_{IF}
\]

where \( f_{IF} \) is the amount that the local oscillator is above the carrier frequency selected. In most AM broadcast band radio receivers \( f_{IF} = 455 \text{ kHz} \), and in most FM receivers \( f_{IF} = 10.7 \text{ MHz} \).

With the latest all digitally tuned radio receivers the basic techniques discussed above still apply. The RF amplifier can be tuned using an LC tank circuit with the capacitance component being a varactor diode which provides a voltage controlled capacitance, and hence a voltage controlled resonant frequency. The local oscillator frequency can be obtained from a digitally programmable frequency synthesizer. The tuning voltage which drives the voltage controlled oscillator of the frequency synthesizer can also be used to tune the RF amplifier. In any case a digital word which corresponds to the desired station frequency can be used to set both the local oscillator and the RF amplifier center frequency.

2.4.1 Frequency Translation

Preliminary Analysis

1. Calculate the frequencies at the output of the first mixer (multiplier), at the input of the 455 kHz IF filter, and at the output of the 455 kHz IF filter. Here assume that \( f_{LO} \) is the value of (2.32).
2. Calculate the range of frequencies over which the local oscillator must be capable of tuning if \( f_{IF} = 455 \text{kHz} \) and \( f_c \) may vary from 550 kHz to 1600 kHz, as in standard AM broadcast.

3. In this portion of the prelab you will perform a more detailed analysis of the signal flow from transmitter through receiver. In Figure 2.26 the signal points of interest are labeled with circled numbers 1–5. For each signal point obtain (i) Time domain expressions with amplitudes, (ii) Frequency domain expressions with amplitudes, and (iii) frequency domain sketches using numerical values for your frequencies. Assume that the modulation index is \( a = 1/2 \), \( f_c = 500 \text{kHz} \), \( m(t) \) is sinusoidal with frequency \( f_m = 5 \text{kHz} \). What you need to include in your lab report is the following:

(a) Verify the results (answers) given below for points 1 and 2.

(b) Perform a similar analysis for points 3 and 4. Point 5 should be of the same form as \( m(t) \), i.e., \( \hat{m}(t) \approx m(t) \).

- Note that \( W \) is the amplitude of the LO and not a bandwidth.
- Assume \( f_{LO} = f_c + f_{IF} = 500 + 455 = 955 \text{kHz} \). Use these results as a theoretical basis for comparing observations to theory later in the lab experiment.
- **Do not** at this point actually measure amplitudes in the lab.
- The amplitudes in this analysis are only a mental exercise; we are most concerned with the presence or absence of a signal at a given frequency.

---

Figure 2.26: Superheterodyne Receiver Circuit
2.4. AM SUPERHETERODYNE RECEIVERS

Solutions for 1 and 2
We use the fact that

\[ A \cos(2\pi f_o t) \Leftrightarrow \frac{1}{2} A \delta(f - f_o) + \frac{1}{2} A \delta(f + f_o) \]

\[ x(t) \cos(2\pi f_o t) \Leftrightarrow \frac{1}{2} X(f - f_o) + \frac{1}{2} X(f + f_o) \]

\[ \cos(u) \cos(v) = \frac{1}{2} [ \cos(u + v) + \cos(u - v) ] \]

in the solutions.

For point 1 it follows that (see Figure 2.27):

\[ m(t) = V \cos(2\pi f_m t), \quad 5\text{kHz} \]

\[ M(f) = \frac{1}{2} V \delta(f - f_m) + \frac{1}{2} V \delta(f + f_m) \]

![Figure 2.27: Sinusoidal message spectrum for \(f_m = 5\text{ kHz}\)](image)

For point 2 it follows that (see Figure 2.28):

\[ s(t) = \left[ A + m(t) \right] B \cos(2\pi f_c t), \quad f_c = 500\text{kHz} \]

Now, when the modulation index is \(a = 1/2\) and \(m(t) = V \cos(2\pi f_m t)\),

\[ s(t) = 2V \left[ 1 + \frac{1}{2} \cos(2\pi f_m t) \right] B \cos(2\pi f_c t) \]

\[ = 2VB \cos(2\pi f_c t) + \frac{1}{2} VB \left\{ \cos[2\pi (f_c + f_m) t] + \cos[2\pi (f_c - f_m) t] \right\} \]

and

\[ S(f) = VB \left\{ \delta(f - f_c) + \delta(f + f_c) \right\} + \frac{1}{4} VB \left\{ \delta[f - (f_c + f_m)] + \delta[f - (f_c - f_m)] + \delta[f + (f_c + f_m)] + \delta[f + (f_c - f_m)] \right\} \]
Laboratory Exercises

To simulate a superheterodyne receiver, you will need to construct the circuit shown in Figure 2.25. Note that the carrier source, multiplier, and 5000 Hz sinusoidal message source can be replaced by a single Agilent 33120A. Note the Agilent 33250A may be used too, but then the 33120A will be needed for the receiver local oscillator.

1. Adjust the local oscillator (LO) sine wave source to about 500 mv peak-to-peak with a frequency of 955 kHz. With the Agilent synthesized sources getting the precise LO frequency needed to down-convert the RF signal, should be easy.

2. Adjust the AM modulator and/or function generators to obtain a 50% modulated 500 kHz carrier at a level that will not cause distortion in the mixer output. Be sure to balance the receiver mixer.

3. Adjust the LO frequency slightly to obtain a maximum output from the IF filter. When you have the maximum, accurately measure the frequency of each of the signal sources into the mixer and calculate the center frequency of the filter.

   - When measuring the frequency of the modulated carrier, measure it with 50% modulation applied to it. Use the frequency counter to measure the carrier frequency. Do you know why the frequency counter can accurately measure a 50% modulated waveform?

4. With the oscilloscope triggered by the 5000 Hz modulation source, observe the input and output of the IF filter. Note that the envelope, or modulation, is the same for either signal. The output, however, should be a clean and undistorted modulated carrier at the IF. With the output of the filter at a maximum, adjust the frequency control on the 5000 Hz source until you can clearly see the IF carrier waveform and thereby verify that the double frequency component is suppressed by the filter. Note: the modulation index of the AM signal at the output of the IF filter will change depending upon how the carrier and/or local oscillator frequencies are adjusted. This is due to the fact that the 5000 Hz message sidebands get attenuated due to the narrow bandwidth of the IF filter.

5. Using a spectrum analyzer connected to the input and then to the output of the mixer and filter, identify all the frequencies components of the superheterodyne system you have constructed. Make a sketch of the spectrum at each of the locations you check. Note: the 50-ohm input impedance of the Agilent 4395A analyzer will load your circuit excessively. Use the high impedance active probe to solve this problem.
2.4. AM SUPERHETERODYNE RECEIVERS

- Note that if unpredicted frequency spectrum components appear on the analyzer they may be due to unbalanced modulator carrier frequency terms or due to distortion because of overdriving the multiplier. If such terms are present, make an attempt to minimize them by appropriate adjustments of your signal levels and/or balance controls.
- It is recommended that you adjust your carrier frequency to exactly 500 kHz and your message signal frequency to exactly 5 kHz. This will simplify your bookkeeping.
- Carefully examine the specifications of your ceramic IF filter. Is it an SFU or CFU?

6. A good radio receiver should have the same sensitivity for each carrier frequency you tune to on the dial. That is, for a given signal level into the antenna, the receiver output should remain constant no matter what frequency the signal may be. Connect the op-amp/diode RC envelope detector at the IF filter output, using 10 k and 0.01 μf for the RC time constant. Set the sinusoidal message frequency to 500 Hz. Measure the RMS value of the detected 500 Hz waveform. Change the transmitter frequency to 0.6, 0.8, 1.0, and 1.2 MHz, adjusting the LO frequency for maximum output each time, and measure the detected output for each received frequency. Plot this data in dB referenced to the maximum measured sensitivity, versus frequency, so that you obtain a sensitivity curve for your system. Comment on your results.

2.4.2 Receiver Selectivity

The selectivity of a receiver is determined primarily by the IF filter frequency response. In this portion of the experiment you will measure the frequency response of the 455 kHz ceramic IF filter contained in the system box.

Laboratory Exercises

1. Set up the circuit shown in Figure 2.29 making sure that the proper filter input/output impedance levels are maintained. You may need assistance from your lab instructor in determining which filter type, SFU or CFU, the system box at your lab station contains.

2. Using the vector network analyzer Agilent 4395A obtain a plot of the filter gain in db and also the phase response, both versus frequency. Be sure to use a narrow sweep range so that the details of the filter amplitude response can be obtained. Compare the measured filter response with the manufacturers specifications given in Chapter 1 of this manual.

3. Assuming a single filter of this type is used in an AM broadcast receiver, (recall that the minimum channel spacing is 10 kHz) do you feel that the filter skirts are sharp enough to adequately reject adjacent channel stations?

To better understand the selectivity issue consider the VisSim/Comm simulation shown in Figure 2.30, file lab4a.vsm. In the simulation the IF filter characteristics can be changed in a number of ways. Currently the IF filter is a 6th-order Chebyshev Type I, with 1 dB of passband ripple, and 1 dB bandwidth from 452 to 458 kHz. Here adjacent channel signals have been inserted spaced 10 kHz away from the desired signal at 1000 kHz (at 990 and 1010 kHz). Each carrier has 3 kHz single tone modulation at a depth of 50%. Increase the filter
order to 8th-order and see how wide you can make the passband, yet still keep the adjacent channel sidebands down at least 20 dB from the desired sidebands. The upper limit is of course 10 kHz, but is impractical.

4. The IF filter while providing selectivity for the receiver, may also attenuate desired high frequency sidebands of a DSB or AM signal that is to be demodulated. With standard AM Broadcast the message bandwidth is limited to 5 kHz due to the 10 kHz minimum channel spacing. Even if the received message contains information above 5 kHz (high fidelity audio requires about 20 kHz) the IF filter will remove it.

To observe this bandlimiting effect use an FM radio music signal as the message source for the AM modulator with a modulation index of about 50% in the circuit of Figure 2.24. Use a carrier frequency of 1 MHz. Detect and amplify the modulation using the audio amplifier/speaker. Compare the input message signal to the received message signal on an oscilloscope and by listening to it. Note that your AM transmitter does not limit the sidebands to ±5 kHz.

### 2.4.3 Demodulation of Local AM Broadcast Stations

In this final section of the experiment you will use the high gain IF amplifier contained in the system box to build a superheterodyne receiver with enough sensitivity to allow reception of local AM broadcast stations.
Figure 2.30: VisSim/Comm simulation file for AM modulation and superheterodyne receiver
2.4. AM SUPERHETERODYNE RECEIVERS

Laboratory Exercises

1. Construct the superheterodyne receiver circuit shown in Figure 2.31 using the high gain IF amplifier. At the mixer input connect a long wire to serve as an antenna.

2. Tune the local oscillator over the appropriate range of frequencies recalling that broadcast band AM stations are located from 550 kHz to 1600 kHz.
   To find stations initially you may need to have the IF amplifier gain at the maximum setting. If you have a good antenna and tune to a strong station, the IF amplifier will most likely overload and start clipping the output signal. By monitoring the IF amplifier output with the scope you can back the IF gain down to avoid this problem. In a “real” receiver this operation is accomplished with an automatic gain control (AGC) circuit. The AGC circuit would require a feedback connection from the detector output to the gain control pin of the high gain IF amplifier chip (MC 1350).

3. Verify that stations can be tuned in with the LO 455 kHz above or 455 kHz below the desired broadcast station center frequency. If possible log the call letters of all the stations you are able to receive.
2.5 Frequency Modulation and Demodulation

In this experiment frequency modulation (FM) and demodulation will be investigated. Particular emphasis will be placed on the modulation process using a voltage controlled oscillator. The spectrum of an FM signal will also be examined. To demonstrate demodulation of FM a simple slope detection scheme will be used. The simplicity of the slope detection circuitry will become evident along with some of the performance limitations.

2.5.1 FM Frequency Deviation Constant

When the instantaneous frequency of a sinusoidal carrier waveform is proportional to a message, \( m(t) \), it can be expressed as

\[
fi = fc + fd m(t)
\]  

(2.33)

where \( fc \) is the carrier frequency, \( m(t) \) is the modulating signal, and \( fd \) is the frequency deviation constant with units of Hz/volt.

Since frequency is the time derivative of phase, or instantaneous phase is the integral of instantaneous frequency, the FM waveform can be expressed as

\[
x_c(t) = A_c \cos[\theta_c(t)]
\]  

\[
= A_c \cos \left[ 2\pi \left( fc t + fd \int_0^t m(\tau) d\tau \right) \right]
\]  

(2.34)

When \( m(t) = A_{dc} \) = a constant, the instantaneous frequency becomes

\[
fi = \frac{d}{dt}[fc t + fd A_{dc} t]
\]  

\[
= fc + fd A_{dc}
\]  

(2.35)

That is, a dc voltage produces a frequency that is offset from the carrier frequency by \( fd A_{dc} \) Hz.

Laboratory Exercises

Using a power supply, dc voltmeter, and frequency counter, measure the frequency versus dc voltage input for the Agilent 33250A function generator. Of the two synthesized sources, only the Agilent 33250A can be used as a voltage controlled oscillator (VCO). As shown in Figure 2.32, the 33250A is first set for FM modulation, and then an external modulation source, applied via a rear panel connector, is selected. On the front panel select a deviation of 100 kHz. When in external FM mode the deviation value is no longer the peak deviation. Instead it is the peak deviation about the carrier when the modulation input swings to \( \pm 5v \). So for an input of +5 v the generator output should increase in frequency by 100 kHz. For an input of -1 v the generator output should decrease by \( 100/5 \times 1 = 20 \) kHz. Reference all data that you take in the lab to the rear panel input, hereafter referred to as \( v_{mod} \).

1. Using rectangular graph paper and plotting data as it is taken, plot the output frequency versus voltage when \(-5 < A_{dc} < 5\) volts is applied at the \( v_{mod} \) input and the carrier center
Choose External FM

To characterize as VCO apply DC from a power supply use fine voltage adjust.

Figure 2.32: Configuring the Agilent 33250A as a VCO
frequency is 200 kHz using a front panel deviation of 100 kHz. Take data points about 0.5 volts apart as long as your graph is linear. Clearly establish the region of linearity of control by taking more points near the ends as required.

2. Repeat the above for carrier center frequency settings of 100 kHz and 1 MHz, keeping the front panel deviation value set to 100 kHz.

3. Compute $f_d$ for the Agilent 33250A with respect to the $v_{mod}$ input using the data you obtained above for each center frequency. Note that according to Agilent, in all cases we expect $f_d = 100/5 = 20$ kHz/v.

2.5.2 FM Modulation Index - $\beta$

If we let $m(t) = A_m \sin [2\pi f_m t]$ in the earlier equation given for $x_c(t)$, then

$$x_c(t) = A_c \cos \left[ 2\pi f_c t + \frac{A_m f_d}{f_m} \cos(2\pi f_m t) \right]$$

$$= A_c \cos \left[ 2\pi f_c t + \beta \cos(2\pi f_m t) \right]$$

(2.36)

In the above equation, $A_m f_d/f_m = \beta$ is referred to as the modulation index. It is an important parameter for characterizing the FM signal. Note that $\beta$ increases with increasing amplitude of the modulating signal, and decreases with increasing frequency, $f_m$.

The equations given here represent sinusoidal frequency modulation, or tone modulation with frequency $f_m$. The equation is periodic and has a Fourier series. The series is rather complicated, however, and has coefficients which are functions of $\beta$ instead of being fixed constants as in the case, for example, of a square wave. The Fourier series representation of sinusoidal frequency modulation is given by

$$x_c(t) = A_c J_0(\beta) \cos \omega_c t$$

$$+ A_c J_1(\beta) [\cos(\omega_c + \omega_m) t - \cos(\omega_c - \omega_m) t]$$

$$+ A_c J_2(\beta) [\cos(\omega_c + 2\omega_m) t + \cos(\omega_c - 2\omega_m) t]$$

$$+ A_c J_3(\beta) [\cos(\omega_c + 3\omega_m) t - \cos(\omega_c - 3\omega_m) t] \ldots$$

(2.37)

This equation shows that for tone modulation the spectrum consists, theoretically at least, of sidebands spaced $f_m$ apart out to infinite frequency. The function $J_n(\beta)$ is called the n-th order Bessel function, of the first kind, with argument $\beta$. The Bessel functions are graphed and tabulated in most math handbooks, and the value of the function for any given argument, $\beta$, can easily be found.

For $\beta = 0$, $J_0(\beta) = 1$ and all of the remaining coefficients are zero. This says that the only term present in the series in the unmodulated case is the carrier frequency, as it should be. For $\beta$ very small the first pair of sidebands will come and go according to the value of their corresponding coefficients in the equation above.

Laboratory Exercises

1. Connect the output of the Agilent 33250A function generator to a spectrum analyzer. Use a carrier frequency of about 1 MHz and center the carrier signal on the analyzer screen. Use the Linear vertical scale and a frequency scan of 10 kHz per division.
2. Connect a sinusoidal modulating signal (Agilent 33120A) to the $v_{mod}$ input of the 33250A with a frequency of 10 kHz. Start with the modulating signal at zero amplitude and **slowly** increase the level. Watch the 33250A output on a scope to observe the frequency modulated waveform and also observe the frequency spectrum. As the input level is increased, one pair of sidebands and then a second and a third pair will appear. Reduce the input until only the first pair is present. Then find the point where the second pair is just barely begins to be noticeable. Take data to calculate the value of $\beta$ at this point. The condition where only one sideband of the modulating frequencies is present is known as **narrow-band** FM. The maximum modulation index for sinusoidal narrow-band FM is usually assumed to be around $\beta = 0.5$ or less. Would you agree?

3. Increase the modulation amplitude slowly, observing the FM waveform and noting the appearance of several additional sidebands on the analyzer screen. Use the 10 dB per division vertical scale. At some point, after three or four sidebands have appeared, the carrier frequency line will begin to decrease in amplitude. Adjust the modulation amplitude until the carrier term is gone. Calculate $\beta$ for this condition and compare your experimental value with theory. Note that the zeros of the $J_n(\beta)$ Bessel functions are tabulated in mathematical handbooks and communication theory texts such as Ziemer and Tranter\(^3\). A short table of zeros is given below

<table>
<thead>
<tr>
<th>$n$</th>
<th>$J_n(x) = 0$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>2.4048 5.5201 8.6537 11.7915 14.9309</td>
</tr>
<tr>
<td>1</td>
<td>3.8317 7.0156 10.1735 13.3237 16.4706</td>
</tr>
<tr>
<td>2</td>
<td>5.1356 8.4172 11.6198 14.7960 17.9598</td>
</tr>
<tr>
<td>4</td>
<td>7.5883 11.0647 14.3725 17.6160 20.8269</td>
</tr>
<tr>
<td>5</td>
<td>8.7715 12.3386 15.7002 18.9801 22.2178</td>
</tr>
<tr>
<td>6</td>
<td>9.9361 13.5893 17.0038 20.3208 23.5861</td>
</tr>
<tr>
<td>8</td>
<td>12.2251 16.0378 19.5545 22.9452 26.2668</td>
</tr>
</tbody>
</table>

4. Now use one-half the modulation frequency of that used above and, by adjusting the input signal amplitude, duplicate the conditions observed in 3. How do the amplitudes compare for these two cases? Does this agree with the equation for $\beta$?

5. Return to the frequency and amplitude of 3. Increase the input signal level slowly and note the signal amplitudes for which $J_1(\beta)$, $J_2(\beta)$, etc., go to zero. At what value of $\beta$ does the carrier term go to zero a second time? Compare your results with theory. Additionally compare your results to simulation experiments run using the VisSim/Comm file lab5a.vsm depicted in Figure 2.33.

6. The parameter $\beta$ is sometimes written as $\beta = \Delta f / f_m$. That is, $f_d A_m = \Delta f$, and $\Delta f$ is called the frequency deviation. It is the maximum instantaneous frequency or the peak swing in carrier frequency from its unmodulated value. The bandwidth occupied by an FM

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Figure 2.33: VisSim/Comm simulation file for sinusoidal and squarewave FM
2.5. FREQUENCY MODULATION AND DEMODULATION

signal is related to the peak deviation, but not in rigorous fashion. Set the peak deviation, \( f_d A_m = \Delta f \), at a fixed value that gives sidebands out to about 100 kHz on each side of the carrier (use 50 kHz per division on the analyzer).

7. Decrease the modulating frequency without changing \( \Delta f \), and notice the effect on the spectrum. Measure the bandwidth at several different modulating frequencies. If you are using the 10dB per division vertical scale you can define spectral bandwidth in terms of the frequencies for which the spectrum is say 10 or 20dB down from its peak value. Calculate the relationship between \( \Delta f \) and bandwidth for each. Carson’s rule for FM by a sinusoidal signal of frequency \( f_m \) states that the bandwidth is approximately

\[
BW = 2 f_m (\beta + 1) = 2(\Delta f + f_m)
\]

Would you agree with this approximation?

2.5.3 FM with Other than Sinusoidal Signals

The Fourier series (or transform) for an FM waveform is mathematically tractable for only a few special cases, the sinusoidal case being one of them. For signals which are more complicated, the detailed structure of the spectrum cannot be analyzed. Only a few rules relating modulating frequency, bandwidth, and peak deviation can be used to describe the frequency domain representation of FM for the general case.

Suppose \( m(t) \) is a zero average value square wave of amplitude \( \pm A_m \). Then the instantaneous frequency of the resulting FM signal jumps from \((f_c - f_d A_m)\) to \((f_c + f_d A_m)\) or from \((f_c - \Delta f)\) to \((f_c + \Delta f)\). Mathematically we can write this as

\[
x_c(t) = \sum_{n=-\infty}^{\infty} p(t - n T_m)
\]

where

\[
p(t) = A_c \left\{ \Pi \left( \frac{t - T_m/4}{T_m/2} \right) \cos[2\pi(f_c + \Delta f)t] \right. \\
+ \left. \Pi \left( \frac{t - 3T_m/4}{T_m/2} \right) \cos[2\pi(f_c - \Delta f)t] \right\}
\]

For this special case the power spectral density of \( x_c(t) \) is relatively easy to obtain. Clearly \( S_x(f) \) will consist of delta functions spaced at multiples of \( 1/T_m \). The envelope of \( S_x(f) \) is proportional to \( |P(f)|^2 \).

Laboratory Exercises

1. Using the Agilent 33250A function generator with a carrier frequency of 1 MHz, apply a square wave signal of 10 kHz to the \( v_{mod} \) input.

2. Observe the modulated waveform on a scope triggered by the 10 kHz signal. Increase the frequency deviation of the Agilent 33250A function generator from 100 to 800 kHz. Try to
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adjust the modulating signal amplitude so that the two output frequencies are clearly evident. This should be observed at an input signal level of about 0.5v peak-to-peak, and at several more values up to as high as about 5.5v peak-to-peak. Try using a sweep rate of 20 μs/div on the oscilloscope.

3. Observe the spectrum on the analyzer as Δf is increased. Use the analyzer with zero frequency in the center and with 500kHz per division so that both positive and “negative” frequencies can be observed.

4. Slowly increase and decrease the modulation amplitude. Notice that at high levels the spectrum seems to cross the zero frequency line.

5. Return the 33250A frequency deviation back to 100 kHz as in Section 2.5.2. Set the amplitude for a ±100 kHz bandwidth and decrease the modulation frequency without changing Δf. Does Carson’s rule seem to hold for this signal?

6. Use an FM radio for a voice/music signal to modulate the 33250A generator. Set the bandwidth at about ±100 kHz for this “random” signal and observe the spectrum. Manually increase the analyzer sweep speed so that you can more clearly see the constant changes in the spectral picture for this general case.

2.5.4 FM Demodulation using Slope Detection

To recover the information contained in an FM signal requires obtaining the signals instantaneous frequency. For the signal

\[ x_c(t) = A_c \cos \left[ \omega_c t + kf \int^t m(\alpha) d\alpha \right] \]  

the instantaneous radian frequency is

\[ \omega_i = \omega_c + kf m(t) \]  

where \( k_f = 2\pi f_d \). A network which has a magnitude response of the form \( |H(f)| = a(f - f_c) + b \) in the vicinity of the carrier frequency \( f_c \), will convert \( x_c(t) \) into an AM type signal with an envelope proportional to the instantaneous frequency. Envelope detection can then be used to obtain \( m(t) \). Mathematically the easiest example is the ideal differentiator which has transfer function \( j2\pi f \). The differentiator output is

\[ x'_c(t) = A_c [\omega_c + kf m(t)] \sin \left[ \omega_c t + kf \int^t m(\alpha) d\alpha \right] \]  

where the envelope \( A_c [\omega_c + kf m(t)] \) is clearly proportional to \( m(t) \) since \( \omega_c \) is fixed.

Practical FM demodulator implementations are shown in Figures 2.34 and 2.35. The first approach uses the transition band of a highpass filter to obtain an \( |H(f)| \) function of the desired form. Since the slope of \( |H(f)| \) is used to recover \( m(t) \), this approach is known as slope detection. In Figure 2.34 a bandpass filter with \( f_c \) set slightly above or below the filter center frequency is used. The problem with both of these implementations is that the assumption that \( |H(f)| = \)
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Figure 2.34: Highpass Filter Slope Detector

Figure 2.35: Bandpass Filter Slope Detector
2.5. FREQUENCY MODULATION AND DEMODULATION

Figure 2.36: Balanced Discriminator

![Balanced Discriminator Diagram]

Figure 2.37: Slope Detector

![Slope Detector Diagram]

$a(f - f_c) + b$ is valid only over a narrow band of frequencies. A more accurate model for $|H(f)|$ near $f_c$ would be to expand $|H(f)|$ in a Taylor series so that

$$|H(f)| = \sum_{n=0}^{\infty} a_n (f - f_c)^n$$

(2.44)

The second order and above terms will cause distortion in the demodulated output. The degree of the distortion will depend upon the peak frequency deviation of the FM signal. The balanced discriminator shown in Figure 2.36 provides improved performance due to its wider linear response region. The phase-locked-loop, which is a topic for a future experiment, can also be used to demodulate FM.

**Laboratory Exercises**

1. Using the highpass filter and amplifier/envelope detector circuits contained in the system box, set-up the slope detection system shown in Figure 2.37.

2. As an FM source use the Agilent 33250A in the FM modulation mode. Use a sinusoidal message at 1 kHz keeping the FM deviation small initially to avoid distortion. Set the carrier frequency to about 50 kHz and observe the HPF output on the oscilloscope as the filter cutoff frequency is adjusted. Obtain an AM type signal at the filter output. You may use the internal sinusoidal modulation capabilities of the 33250 for this portion if you wish.
3. For a fixed FM deviation and fixed filter cutoff frequency, determine experimentally if there is a carrier frequency which minimizes the harmonic distortion observed in the demodulated output. Note that the Agilent 4395A can perform a peak search, which can be helpful in finding the relevant harmonics. The harmonic distortion is the percent of power in the harmonics \( (n \geq 2) \) divided by the total signal power. Having found an “optimum” carrier and cutoff frequency combination comment on how the distortion changes when the FM deviation is altered. Check to see if the tone modulation frequency has any significant effect on the distortion. Comment on the above observations from a theoretical standpoint.

4. Test the demodulation system using an FM radio as a voice/music message source. Note that you will now have to apply external modulation via the rear, \( v_{\text{mod}} \) input. Be sure to match the amplitude level from the radio to an appropriate deviation sensitivity. Recall that \( f_d = \text{generator deviation}/5 \text{ Hz}/v \).
2.6 Second Order Phase-Lock Loops

The purpose of this laboratory exercise is to design, build, and experimentally characterize a second order phase-lock loop (PLL). Two types of loop filters will be investigated, the lead-lag filter and the integrator with lead compensation. The PLL will be designed to operate at a specific carrier center frequency, \( f_0 \), loop natural frequency, \( \omega_n \), damping factor, \( \zeta \), and hold-in range, \( \Delta f_H \). Note that the hold-in range is the maximum amount the input frequency can deviate from the quiescent VCO frequency and have the loop remain in lock. This term assumes the PLL is initially in a locked state.

The PLL will be constructed using for the most part pre-assembled circuit function blocks mounted on an RF bread-board. The loop filter will use a bi-fet op-amp with appropriate resistor and capacitor elements determined from the design equations. Before the actual PLL design begins, the phase detector and the VCO must be characterized. For this experiment the phase detector is a surface mount packaged double-balanced mixer using transformers and a diode ring (Mini-Circuits ADEX-10L) and the VCO is a surface mount packaged part (Mini-Circuits JTOS-75). The VCO output is buffered through a surface mount two-way 50 ohm power splitter (Mini-Circuits ADP-2-1W) so that a 50 ohm test port is always available to drive a spectrum analyzer/scope/frequency counter. The reference signal \( s(t) \) will be obtained from an 80 MHz Function/Arbitrary Waveform Generator, (Agilent 33250A).

The system requirements are to design a second order PLL to initially have:

- Nominal carrier center frequency \( f_0 = 50.000 \) MHz
- Loop Hold-In Range \( \Delta f_H = 5 \) MHz
- Natural Frequency \( \omega_n = 5 \times 10^4 \) rad/sec
- Damping Factor \( \zeta = 1/\sqrt{2} \)

Later in the experiment you may want to change some of these specifications, if time permits.

The loop filters considered will be a lead-lag filter with finite DC gain and an integrator with lead correction filter having in theory infinite DC gain. The close-loop parameters \( \omega_n \) and \( \zeta \) will be experimentally measured, and step response measurements will be taken.

If time permits simple nonlinear acquisition measurements will be taken. Additionally, the phase noise will also be measured and compared with a behavioral level model implemented using Spice.

2.6.1 PLL Subsystem Characterization

For the design of a PLL from scratch you would first consider the system requirements before selecting any of the loop subsystems (i.e. phase detector, loop amplifier/filter, and VCO). For this experiment you do not have this kind of freedom. The laboratory has been designed to work with the available equipment, with certain compromises being made. A block diagram of the PLL showing the various laboratory subsystems is given in Figure 2.38. Everything to build the complete PLL, except the Agilent 33250A reference source, is mounted on the test board. A ±12v power supply is also needed to power the active circuits on the board. The VCO needs 12v and
the op-amp requires ±12v. A corresponding photograph of just the PLL module itself is shown in Figure 2.39. The relevant elements are noted in this photo. Additional circuitry on the board

![Laboratory PLL block diagram](image)

Figure 2.38: Laboratory PLL block diagram.

not immediately evident from the block diagram and photo are an RC lowpass filter consisting of a 51 ohm resistor and a 0.01 μf chip capacitor, and power decoupling filters on the input power jacks to the board and additionally on the VCO input power pin. These two circuits are detailed in Figure 2.40 One other circuit item mounted on the board next to the op-amp is a 5 kΩ multi-turn potentiometer. This will be used to form an adjustable bias voltage which can be used create a level-shifter for setting the VCO quiescent frequency to 50 MHz.

To reduce Figure 2.38 to the equivalent linear PLL block diagram we first must characterize the Mini-Circuits JTOS-75 VCO to obtain $K_v$, and then characterize the phase detector to obtain $K_p$.

### 2.6.2 Laboratory Exercises – Part I

#### VCO Characterization

Detailed JTOS-75 VCO data sheets can be found in the appendix. We wish to characterize the VCO in the vicinity of 50 MHz to determine an approximate value for $K_v$. The VCO out port on the board can be used to drive a frequency counter. This port is the result of feeding the VCO output through a 2-way equal gain, equal phase, power splitter (Mini-Circuits ADP-2-1W).

From the data sheet the VCO produces 50 MHz when the tuning voltage is at about 7 v. To actually move the VCO frequency you will need to connect a bias supply to the VCO control input. This can be supplied from an additional supply or by configuring an op-amp as a level-shifter and using the multi-turn pot already on the board. A possible configuration for an op-amp level-shifter bias supply circuit is shown in Figure 2.41.

- Obtain frequency as a function of control voltage data so you can plot $f_0$ versus voltage from 40 to 60 MHz. This will not be a perfectly straight line. The slope at 50 MHz however is a good estimate for $K_v$. Remember later during loop calculations to convert $K_v$ to rad/s/v.
2.6. SECOND ORDER PHASE-LOCK LOOPS

![PLL module photograph.](image)

Figure 2.39: PLL module photograph.

![Diagram of PLL components and connections.](image)

Figure 2.40: Details on phase detector lowpass and power supply decoupling circuits.
A second plot to make from the recorded data is a plot of $K_v$ in MHz/v as a function of center frequency. A plot similar to this is in the data sheet, so you have something to compare to for your particular VCO.

The fact that VCO sensitivity is not constant gives you some idea of how loop parameters will change should you want to move the center frequency of the loop. For future reference, note that the VCO data sheet indicates that the modulation bandwidth of the VCO is 125 kHz, this is related to the VCO spurious time constant.

**Phase Detector Characterization**

The phase detector consists of a Mini-Circuits ADEX-10L double-balanced mixer (DBM), followed by a lowpass filter to remove the double frequency terms. Note that in the analysis of the complete PLL this lowpass filter constitutes a spurious time constant. Since the mixer output impedance is approximately 50 ohms, the R value in the lowpass filter is actually about 100 ohms. The specifications for the ADEX-10L mixer can be found in the Appendix.

A DBM phase detector is very similar to an ideal multiplier or sinusoidal phase detector. The phase detector constant $K_p$ is a function of the signal amplitudes at both the RF (reference source) port and the LO (VCO) port. For this experiment the RF drive level will be -5dBm. The Agilent 33250A, which serves as the reference signal source, can be easily configured to this output power level via front panel settings.

To experimentally determine $K_p$ use the test set-up shown in Figure 2.42. Two sinusoidal signals, each at approximately 50 MHz, are input to the DBM RF and LO ports. The input signals are of the form

\[
v_{RF}(t) = A \sin(2\pi f_0 t + \theta)
\]

\[
v_{LO}(t) = B \cos(2\pi f_0 t + 2\pi \Delta f t + \hat{\theta})
\]
2.6. SECOND ORDER PHASE-LOCK LOOPS

Figure 2.42: Test set-up for measuring phase detector gain.

where $\Delta f$ is a small frequency difference between the two signals. Using a sinusoidal phase detector model the phase detector output, assuming the RC lowpass filter removes the double frequency term is

$$v(t) = \frac{AB}{2} K_m \sin(2\pi \Delta f t + \theta - \hat{\theta})$$

$$= K_P \sin(2\pi \Delta f t + \theta - \hat{\theta})$$

(2.47)

where $K_m$ is the mixer multiplication constant and

$$K_P = \frac{AB}{2} K_m \text{ v/ rad}$$

(2.48)

is defined to be the phase detector gain. From (3) it follows that $K_P$ can be experimentally determined by observing the phase detector output for small $\Delta f$. Note that if $\Delta f$ is not small, then the phase detector lowpass filter will also significantly attenuate the frequency difference term as given by (3). When the phase detector is placed in the PLL and the loop is locked, then of course $\Delta f$ is essentially zero.

For proper operation of a DBM the LO drive level is typically at least 10 dB higher than the RF power level. For the ADEX-10L the recommended LO drive level is 4 dBm to insure minimal conversion loss (see the mixer specs in the appendix). The VCO output is nominally +8 dBm, but the 2-way splitter drops 3 dB, so the LO input to the mixer should be about +5 dBm. It is important to note here that for the 3 dB splitter loss to hold, the VCO test port must be kept terminated into a 50 ohm load, i.e., some test instrument or if need be a 50 ohm BNC terminator, which can be found in the lab. An RF level of -5 dBm is 10 dB below the LO level, and according to the data sheet the conversion loss at 50 MHz should be about 7.2 dB (ideally 6 dB due to the 1/2 factor of the trig identity of (3)).

- Using the test set-up of Figure 2.42 apply 50 MHz sinusoids to both the RF and LO ports of the ADEX-10L mixer (phase detector). The LO power level is fixed at about +5 dBm assuming the VCO test port is properly terminated. Verify the VCO test port power using the spectrum analyzer to so you know approximately what the true LO drive level really is.
2.6. SECOND ORDER PHASE-LOCK LOOPS

- Take measurements to determine $K_p$ with the RF level at -15, -10, and -5 dBm. Comment on the linearity of $K_p$ in response to changes in the RF power level. You will need to make $\Delta f$ as small as possible to insure that the lowpass filter does not introduce error into your measurement. The waveform that you see on the oscilloscope will look like a frequency modulated sinewave. By increasing $\Delta f$ you will see the envelope of the output signal shrink due to the lowpass filtering action.

It will be hard to make the difference frequency to stand still, since the VCO wants to drift when in an open-loop configuration. Just changing the VCO test port load termination will cause some frequency pulling.

2.6.3 Second Order PLL with Lead-Lag Loop Filter

Now that the phase detector and VCO gains have been determined, the design of the loop filter may proceed. In this section of the experiment an active lead-lag loop filter with transfer function

$$F(s) = \frac{1 + sR_2}{1 + sR_1}$$

will be implemented using the op-amp circuit shown in Figure 2.43. Since the VCO must be biased to the specified 50 MHz center frequency, it is convenient to incorporate a level-shifter into the loop filter. One possible configuration is shown in Figure 2.44.

![Figure 2.43: Basic lead-lag active loop filter.](image)

2.6.4 Laboratory Exercises – Part II

The loop filter components $R_1$, $R_2$, $R_3$, and $C$ must be determined such that the system requirements are satisfied. Designing for a specific natural frequency, $\omega_n$, and damping factor, $\zeta$, follows directly from the PLL linear model equations. The loop hold-in range can be determined from the phase detector characteristic and the dc loop gain. Since the phase detector has a maximum output of $K_p \sin(\pi/2) = K_p$, the hold-in range of the PLL is equal to the dc loop gain $\pm K_p F(0) K_v = \pm K F(0)$ rad/sec. This is the maximum offset frequency that the VCO can be tuned to, with respect to the VCO quiescent frequency, and still remain locked. Note that the loop filter of Figure 2.43 has a dc gain magnitude of $R_2/R_1$. 

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2.6. SECOND ORDER PHASE-LOCK LOOPS

*Figure 2.44: Lead-lag active loop filter including a level shifter for VCO center frequency bias.*

- Comment on the significance with respect to second order PLL design, of the spurious constant due to the lowpass filter in the phase detector and spurious time constant of the VCO.

- Determine suitable component values for $R_1$, $R_2$, $R_3$, and $C$ with an RF power level of -10 dBm to achieve $\omega_n = 5 \times 10^4$ rad/s, $\zeta = 1/\sqrt{2}$, and $\Delta f_H = \pm 5$ MHz. Since a wide range of capacitor values is not available in the lab, it is suggested that you initially choose a convenient value for $C$.

- Construct the loop filter using an TLO-82 (LF353), or similar 8-pin DIP dual op-amp as found on the experiment module. A description of the TLO-82 and its pin-out are given in Figure 2.45. More detailed TLO-82 specifications can be found in the appendix.

- Place the loop filter into the PLL as shown in Figure 2.38. Initially you may want to leave the connection from the phase detector output to the loop filter input open. Do connect the loop filter with level-shifter output up to the VCO input. This will allow you to adjust the VCO quiescent to approximately 50 MHz before closing the loop. When the loop is closed, the quiescent frequency will shift slightly due to a small offset voltage in the mixer (phase detector) output.

- Now close the loop by connecting the phase detector output to the loop filter input. Verify phase lock by observing the VCO output on the Agilent 4395A spectrum/vector network analyzer, or HP 4195A spectrum/vector network analyzer. The spectrum analyzer can be directly connected to the VCO output using the RF power divider output port supplied on the experiment module. You may have to adjust the level shifter offset slightly to bring the loop into lock. When in lock discrete frequency sidebands present on the VCO output should disappear.

- Experimentally determine the loop hold-in range, $\Delta f_H$, by slowly adjusting the frequency of the input signal (RF port) above and below 50 MHz (the loop must be locked to begin with). Record the upper and lower frequencies, if possible using a frequency counter, where the PLL just breaks lock. The difference between these upper and lower lock frequencies is by definition **twice** the hold-in range, $\Delta f_H$. 
description

The TL08x JFET-input operational amplifier family is designed to offer a wider selection than any previously
developed operational amplifier family. Each of these JFET-input operational amplifiers incorporates
well-matched, high-voltage JFET and bipolar transistors in a monolithic integrated circuit. The devices feature
high slew rates, low input bias and offset currents, and low offset voltage temperature coefficient. Offset
adjustment and external compensation options are available within the TL08x family.

The C-suffix devices are characterized for operation from 0°C to 70°C. The I-suffix devices are characterized
for operation from –40°C to 85°C. The Q-suffix devices are characterized for operation from –40°C to 125°C.
The M-suffix devices are characterized for operation over the full military temperature range of –55°C to 125°C.

symbols

Figure 2.45: TLO-82 op-amp description and pin-out.
2.6. SECOND ORDER PHASE-LOCK LOOPS

- Experimentally determine the pull-in range, $\Delta f_p$, by increasing/decreasing the input frequency until the loop breaks lock. Then slowly move the input frequency in the other direction until the loop just pulls in. When done from both directions you will be able to determine $2\Delta f_p$, and hence $f_p$. You should find that $\Delta f_p < \Delta f_H$.

The next exercise will be to experimentally determine the second order loop parameters $\omega_n$ and $\zeta$. The test set-up is shown in Figure 2.46. The approach used here is to apply sinusoidal frequency modulation to the input signal of radian frequency $\Omega$ and peak frequency deviation $\Delta \omega_s = 2\pi \Delta f_s$. Thus the PLL input signal is

$$v_{RF}(t) = A \sin \left[ 2\pi f_0 t + \frac{\Delta \omega_s}{\Omega} \sin(\Omega t) \right]$$ (2.50)

Under steady state conditions the loop error signal $\phi(t)$ is of the form

$$\phi(t) = m_x \sin(\Omega t + \phi_x)$$ (2.51)

where for a high gain ($K \gg \omega_n$) second order loop of the type used here ($N = 1$)

$$m_x \approx \Delta \omega_s \frac{\Omega}{\sqrt{\omega_n^2 - \Omega^2}^2 + 4\zeta^2 \omega_n^2 \Omega^2}$$ (2.52)

Figure 2.46: Test set-up for determining second order parameters.
We also know that \( m_x \) is maximized when \( \Omega = \omega_n \) and the maximum value is given by

\[
\max_{\Omega} m_x = \frac{\Delta \omega_s}{2 \xi \omega_n} \tag{2.53}
\]

If we define \( \Omega' \) and \( \Omega'' \) as the upper and lower 3 dB frequencies with respect to the maximum value of \( m_x \), then it can be shown that

\[
\omega_n^2 = \Omega' \Omega'' \tag{2.54}
\]
\[
2 \xi \omega_n = \Omega' - \Omega'' \tag{2.55}
\]

Thus by experimentally determining \( \Omega' \) and \( \Omega'' \) using a frequency counter and an AC voltmeter connected to the phase detector output, as shown in Figure 2.46, \( \omega_n \) and \( \xi \) can be determined.

- Using the test set-up shown in Figure 2.46 experimentally determine \( \omega_n \) and \( \xi \) of the PLL. Make sure that the peak phase error which occurs when \( \Omega = \omega_n \), does not significantly violate the loop linearity assumptions (e.g., \( \max m_x < \pi / 6 \)). When making this measurement you will notice considerable phase jitter on the phase error signal. This is due to the phase noise of both the input signal function generator and the VCO function generator. The oscillator phase noise creates a source of error that can be seen as a noise voltage bias on the AC voltmeter. The double frequency term at 100 MHz is also present, but can be rejected on the scope display by selecting HF reject as the input filtering mode.

- Turning away from the hardware, construct a VisSim/Comm simulation of the above PLL in baseband form, it does not need to be complex baseband as the nonlinear baseband model will suffice. Using the design values from the hardware PLL just studied, verify the performance of the simulation in terms of hold-in range, pull-in range, and closed loop sinusoidal FM response as a means for experimentally determining (verifying) \( \omega_n \) and \( \xi \).

### 2.6.5 Second Order PLL with Integrator/Lead Compensation Loop Filter

In this section of the experiment the PLL will be redesigned using a loop filter which consists of an integrator with phase lead compensation. The loop filter transfer function is given by

\[
F(s) = \frac{1 + s \tau_2}{s \tau_1} \tag{2.56}
\]

and the op-amp implementation of this filter is as shown in Figure 2.47. From a design standpoint the major difference between this filter and the lead lag filter is that the integrator in the loop filter makes the hold-in range theoretically infinite (i.e. \( |KF(0)| = \infty \)). In practice the hold-in range is only limited by the tuning range of the VCO.

### 2.6.6 Laboratory Exercises – Part III

The loop filter components \( R_1, R_2, \) and \( C \) must be determined such that the system requirements are satisfied. Designing for a specific natural frequency, \( \omega_n \), and damping factor, \( \xi \), follows directly from the PLL linear model equations.
2.6. SECOND ORDER PHASE-LOCK LOOPS

Determine suitable component values for $R_1$, $R_2$, and $C$ for an RF power level of -10 dBm to achieve $\omega_n = 10^3$ rad/s, and $\zeta = 1/\sqrt{2}$. What is $\Delta f_H$? Again it is suggested that you initially choose a convenient value for $C$.

Modify the original lead lag loop filter circuit to reflect the new loop filter component values.

Close the loop by connecting the level shifter output to the VCO control signal input. Verify phase lock by observing the VCO output on the HP 853/8557A spectrum analyzer (if possible use the HP 4195A). A potential problem with this loop filter that you may notice is that before the loop locks or perhaps while it is trying to lock, the op-amp integrator output may drift off into saturation. In practice a sweeping circuit may be connected around the loop filter to prevent saturation from occurring. If this is causing problems with your circuit try placing $R_3$ temporarily back into the circuit, then as soon as the loop locks, carefully remove $R_3$. The loop should remain locked.

Experimentally verify the “infinite” hold-in range provided by this loop filter by slowing raising and lowering the frequency of the Agilent 33250A. Is the hold-in range really infinite?

Using the test set-up shown in Figure 2.46 experimentally determine $\omega_n$ and $\zeta$.

2.6.7 Further Investigations

If time permits you may wish to verify additional PLL characteristics. These include transient response, the effects of changing $\omega_n$ and $\zeta$, and additive noise and oscillator phase noise effects on phase jitter.

Transient Response

The easiest transient response to view with the current test configuration is the frequency step response. This can be obtained by setting the frequency modulation function generator to pro-
To observe the entire response the square wave frequency should be less than $10\omega_n/(2\pi)$. Placing the scope on the VCO control voltage is a good place to observe the transient response due to a frequency step. The VCO control voltage, $e(t)$, is proportional to the instantaneous VCO frequency with the addition of the offset. In terms of an isolated rising edge of the applied square wave frequency modulation,

$$e(t) = v_{\text{offset}} + \frac{\Delta f_s}{K_v} \mathcal{L}^{-1}\left\{ \frac{1}{s^2} H(s) \right\}$$

where the units of $K_v$ is here assumed to be Hz/v. For the lead-lag loop filter we know that for $\xi < 1$

$$\mathcal{L}^{-1}\left\{ \frac{1}{s^2} H(s) \right\} = \mathcal{L}^{-1}\left\{ \frac{1}{s^2} \cdot \frac{(2\xi \omega_n - \omega_n^2/K) s + \omega_n^2}{s^2 + 2\xi \omega_n s + \omega_n^2} \right\}, \quad \xi < 1$$

$$= 1 - e^{\xi \omega_n t} \left[ \cos(\omega_n t \sqrt{1 - \xi^2}) - \frac{\xi - \omega_n / K}{\sqrt{1 - \xi^2}} \sin(\omega_n t \sqrt{1 - \xi^2}) \right] u(t)$$

and for $K > \omega_n^2$ we have the same result as for the integrator with lead correction filter,

$$\mathcal{L}^{-1}\left\{ \frac{2\xi \omega_n s + \omega_n^2}{s^2 + 2\xi \omega_n s + \omega_n^2} \right\}, \quad \xi < 1$$

$$= 1 - e^{\xi \omega_n t} \left[ \cos(\omega_n t \sqrt{1 - \xi^2}) - \frac{\xi}{\sqrt{1 - \xi^2}} \sin(\omega_n t \sqrt{1 - \xi^2}) \right] u(t)$$

The expected step response less a DC offset at the VCO input for the second order loops implemented in this lab is shown in Figure 2.48 for various values of $\xi$.

**Changing $\omega_n$ and $\xi$**

Different values of $\omega_n$ and $\xi$ can be considered in terms of the effect on transient response, and also the effect on phase jitter due to oscillator phase noise. Phase jitter is of particular interest here since the reference source, the Agilent 33250A function generator, is not particularly clean. To verify this you may wish to redesign one of the loop filters with $\omega_n$ reduced by a factor of five or ten. Then try to lock the loop and observe changes in the amount of phase jitter that is present. Increasing $\omega_n$ above the original design point should have an opposite effect on phase jitter. With the present low pass filter on the phase detector output $\omega_n$ cannot be increased too much without running into stability problems due to spurious time constant effects.

**Phase Noise Characterization**

The spectrum analyzer can be used to measure the phase noise. To compare with theory, we can create a behavioral level model in Spice (XSpice) that will include phase noise from the input source, VCO, and the noise from the op-amp and the remaining electronic circuitry. Figure 2.49 shows a simulation model useful in modeling the phase noise in the PLL module. Attempting to match theory is another area for further investigation. More information on how to do this effectively may be available in the final release of this lab document.
Figure 2.48: Frequency step response, less a DC offset, as seen at the VCO input for a second order PLL with integrator/lead compensation loop filter.

Figure 2.49: Phase noise modeling.
### 2.6.8 Appendix: Data Sheets

**VCO**

- For Surface Mount Environmental Specifications, please click [here](#).
- Re-flow soldering information is available in "Surface Mount" article.
- Non-hermetic
- Operating Temperature: -55°C to +85°C
- Absolute Max. Supply Voltage ($V_{cc}$) +16V and Tuning Voltage ($V_{tune}$) ±18V
- General Quality Control Procedures and Environmental Specifications are given in Mini-Circuits Guar.
- HI-Rel, MIL description are given in HI-Rel and MIL.
- Prices and Specifications subject to change without notice.

#### JTOS-75

<table>
<thead>
<tr>
<th>Frequency (MHz)</th>
<th>Power Output (dBm)</th>
<th>Tuning Voltage (V)</th>
<th>Phase Noise (dBc/Hz) @ 100 kHz, 1 kHz, 10 kHz</th>
<th>Pulling Mhz</th>
<th>Pushing Mhz</th>
<th>Tuning Sensitivity (MHz/V)</th>
<th>3dB Modulation Bandwidth (kHz)</th>
<th>Power Supply Voltage (V)</th>
<th>Current (mA) Max</th>
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</thead>
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<td>TYP. Min. Max.</td>
<td>TYP.</td>
<td>TYP.</td>
<td>TYP.</td>
<td>TYP.</td>
<td>TYP.</td>
<td>TYP.</td>
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<td>TYP.</td>
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* applies to dual output VCO

#### Pin Connections

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#### Tuning Characteristics

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<th>Tuning Sensitivity</th>
<th>Power Output (dBm)</th>
<th>Harmonics Suppression (dBc)</th>
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#### Case Style - BK377 (inch/mm) weight: 3 grams

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<th>D</th>
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<th>F</th>
<th>G</th>
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Tolerance: ±0.1 ±0.03 ±0.015 inch.
2.6. SECOND ORDER PHASE-LOCK LOOPS

---

**TUNING CHARACTERISTICS - JT08-75**

- **Frequency (MHz)** vs. **Tuning Voltage (V)**
- Frequency increases with tuning voltage.

**TUNING SENSITIVITY - JT08-75**

- **Tuning Sensitivity (MHz/V)** vs. **Tuning Voltage (V)**
- Sensitivity decreases with increasing tuning voltage.

**HARMONICS SUPPRESSION - JT08-75**

- **Harmonics Suppression (dBc)** vs. **Tuning Voltage (V)**
- Suppression improves with tuning voltage for different harmonics.

**SSB PHASE NOISE - JT08-75**

- **SSB Phase Noise (dBc/Hz)** vs. **Carrier Offset (Hz)**
- Noise decreases with increasing carrier offset, with F4 showing the best performance.

---
2.6. SECOND ORDER PHASE-LOCK LOOPS

Mixer (Phase Detector)

**Frequency Mixers**

**LO Power Level 4 dBm**

**Pin Configuration**

<table>
<thead>
<tr>
<th>Port</th>
<th>LO</th>
<th>RF</th>
<th>IF</th>
<th>Gnd Ext</th>
<th>Case Gnd Not Used</th>
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<td>2</td>
<td>1,4,5</td>
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</table>

**Outline Drawing**

**Notes:**
- Absolute maximum power, voltage and current ratings:
  - RF power: 50mW
  - Pin 1+ pin 5 current: 40mA
  - 1 dB CCIRPR. +1 dB typ.
- For Surface Mount Environmental Specifications, please click [here](#).
- Conversion loss increases 0.5 dB when IF is above 150 kHz.
- General Quality Control Procedures and Environmental Specifications are given in [Mini-Circuits Guarantee Qual](#).
- HR and ML description are given in HRPR and ML.
- Prices and Specifications subject to change without notice.

**Electrical Specifications**

**ADEX-10L**

**LO Power Level 4 dBm**

<table>
<thead>
<tr>
<th>Frequency MHz</th>
<th>LO/RF Input</th>
<th>Mid-Band Total Range</th>
<th>L</th>
<th>M</th>
<th>U</th>
<th>L</th>
<th>M</th>
<th>U</th>
</tr>
</thead>
<tbody>
<tr>
<td>10.00-1000/DC-600</td>
<td>8.2</td>
<td>8.8</td>
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**Typical Performance Data**

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<tr>
<th>ADEX-10L</th>
<th>Conversion Loss (dB)</th>
<th>ISOL (dB)</th>
<th>LO Isolation L-R (dB)</th>
<th>LO Isolation L-I (dB)</th>
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<td></td>
<td>GHz</td>
<td>GHz</td>
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<td>100.00</td>
<td>8.85</td>
<td>-75.6</td>
<td>-75.6</td>
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</tbody>
</table>

**Mini-Circuits®**

New Product Information

**ADEX-10L**

**ECE 4670 Communication Laboratory Experiments**
2.6. SECOND ORDER PHASE-LOCK LOOPS

Power Splitter

Notes:

- Aqueous washable. For non-aqueous requirements, LRPS units available in case style QQC130.
- For Surface Mount Environmental Specifications, please click here.
- Re-marking information is available in "Surface Mount" article.
- Internal load dissipation: 0.125 Watt.
- Matched power rating: 2 Watt.
- Non-hemetic.
- General Quality Control Procedures and Environmental Specifications are given in Mint-Circuits Guaran.
- Hi-Rel, MIL description are given in Hi-Rel and MIL.
- Prices and Specifications subject to change without notice.

**ADP-2-1W 2 Way-0°**

<table>
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<tr>
<th>Frequency MHz</th>
<th>Insertion Loss, dB</th>
<th>Phase Unbalance Degrees</th>
<th>Amplitude Unbalance, dB</th>
<th>Power Input, W as Combiner as Splitter</th>
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</tr>
<tr>
<td>L = low range (f&lt;10f₀)</td>
<td>M = mid range (10f₀ to f₀/2)</td>
<td>U = upper range (f₀/2 to f₀)</td>
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<td></td>
</tr>
</tbody>
</table>

**Pin Connections**

Port 1 Port 2 Port 3 Port 4 Port 5 Ext Net Used

**FREQ (MHz)** | **Insertion Loss** | **Amplitude Unbalance** | **Phase Unbalance** | **ISOL (dB)** | **ISOL 1/2 (dB)** | **ISOL 1/2 (dB)** | **ISOL 1/2 (dB)** | **ISOL 1/2 (dB)** | **ISOL 1/2 (dB)** |
<table>
<thead>
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<td>0.00</td>
<td>34.50</td>
<td>0.02</td>
<td>22.000</td>
<td>1.14</td>
<td>1.14</td>
<td>1.14</td>
</tr>
<tr>
<td>23.000</td>
<td>3.30</td>
<td>0.000</td>
<td>0.00</td>
<td>33.48</td>
<td>0.02</td>
<td>23.000</td>
<td>1.14</td>
<td>1.14</td>
<td>1.14</td>
</tr>
<tr>
<td>24.000</td>
<td>3.30</td>
<td>0.000</td>
<td>0.00</td>
<td>32.38</td>
<td>0.02</td>
<td>24.000</td>
<td>1.14</td>
<td>1.14</td>
<td>1.14</td>
</tr>
<tr>
<td>25.000</td>
<td>3.30</td>
<td>0.000</td>
<td>0.00</td>
<td>31.16</td>
<td>0.02</td>
<td>25.000</td>
<td>1.14</td>
<td>1.14</td>
<td>1.14</td>
</tr>
</tbody>
</table>

**Case Style - CD636 (inch, min.)** weight: 0.45 grams.

<table>
<thead>
<tr>
<th>A</th>
<th>B</th>
<th>C</th>
<th>D</th>
<th>E</th>
<th>F</th>
<th>G</th>
<th>H</th>
<th>J</th>
</tr>
</thead>
<tbody>
<tr>
<td>280</td>
<td>.310</td>
<td>.310</td>
<td>.220</td>
<td>.102</td>
<td>.100</td>
<td>.050</td>
<td>.030</td>
<td>.030</td>
</tr>
<tr>
<td>7112</td>
<td>7.874</td>
<td>7.588</td>
<td>2.540</td>
<td>4.115</td>
<td>1.357</td>
<td>2.540</td>
<td>0.762</td>
<td></td>
</tr>
<tr>
<td>K</td>
<td>L</td>
<td>M</td>
<td>N</td>
<td>P</td>
<td>Q</td>
<td>R</td>
<td>S</td>
<td>T</td>
</tr>
<tr>
<td>0.000</td>
<td>1.661</td>
<td>7.620</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Tolerance: x ± .1, y ± .03, z ± .015 inch.
2.6. SECOND ORDER PHASE-LOCK LOOPS

**INSERTION LOSS - ADP-2-1W**

**ISOLATION - ADP-2-1W**
Op-Amp

### Electrical Characteristics, $V_{CC} = \pm 15$ V (unless otherwise noted)

<table>
<thead>
<tr>
<th>PARAMETER</th>
<th>TEST CONDITIONS†</th>
<th>TA</th>
<th>TL081M, TL082M</th>
<th>TL084Q, TL084M</th>
<th>UNIT</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{IO}$ Input offset voltage</td>
<td>$V_O = 0$, $R_S = 50 , \Omega$</td>
<td>25°C</td>
<td>3 6</td>
<td>3 9</td>
<td>mV</td>
</tr>
<tr>
<td>$\alpha_{VIO}$ Temperature coefficient of input offset voltage</td>
<td>$V_O = 0$, $R_S = 50 , \Omega$</td>
<td>Full range</td>
<td>18 18</td>
<td>18 18</td>
<td>µV/°C</td>
</tr>
<tr>
<td>$I_O$ Input offset current‡</td>
<td>$V_O = 0$</td>
<td>25°C</td>
<td>5 100</td>
<td>5 100</td>
<td>pA</td>
</tr>
<tr>
<td>$I_B$ Input bias current‡</td>
<td>$V_O = 0$</td>
<td>125°C</td>
<td>20 20</td>
<td>20 20</td>
<td>nA</td>
</tr>
<tr>
<td>$V_{ICR}$ Common-mode input voltage range</td>
<td></td>
<td>25°C</td>
<td>±12</td>
<td>±12</td>
<td>V</td>
</tr>
<tr>
<td>$V_{OM}$ Maximum peak output voltage swing</td>
<td>$R_L = 10 , k\Omega$</td>
<td>25°C</td>
<td>±12 13.5</td>
<td>±12 13.5</td>
<td>V</td>
</tr>
<tr>
<td>$A_{VD}$ Large-signal differential voltage amplification</td>
<td>$V_O = \pm 10 , V$, $R_L = 2 , k\Omega$</td>
<td>25°C</td>
<td>25 200</td>
<td>25 200</td>
<td>V/mV</td>
</tr>
<tr>
<td>$f_I$ Unity-gain bandwidth</td>
<td></td>
<td>25°C</td>
<td>3 3</td>
<td>3 3</td>
<td>MHz</td>
</tr>
<tr>
<td>$g_{m}$ Common-mode rejection ratio</td>
<td>$V_{IC} = V_{ICR}{\min}$, $V_O = 0$, $R_S = 50 , \Omega$</td>
<td>25°C</td>
<td>80 86</td>
<td>80 86</td>
<td>dB</td>
</tr>
<tr>
<td>$k_{SVR}$ Supply voltage rejection ratio ($\Delta V_{CC} / \Delta V_{Q}$)</td>
<td>$V_{CC} = \pm 15 , V$ to $\pm 9 , V$, $V_O = 0$, $R_S = 50 , \Omega$</td>
<td>25°C</td>
<td>80 86</td>
<td>80 86</td>
<td>dB</td>
</tr>
<tr>
<td>$I_{CC}$ Supply current (per amplifier)</td>
<td>$V_O = 0$, No load</td>
<td>25°C</td>
<td>1.4 2.8</td>
<td>1.4 2.8</td>
<td>mA</td>
</tr>
<tr>
<td>$V_{OY/VOS}$ Crossover attenuation</td>
<td>$A_{VD} = 100$</td>
<td>25°C</td>
<td>120 120</td>
<td>120 120</td>
<td>dB</td>
</tr>
</tbody>
</table>

† All characteristics are measured under open-loop conditions with zero common-mode input voltage unless otherwise specified.
‡ Input bias currents of a FET-input operational amplifier are normal junction reverse currents, which are temperature sensitive as shown in Figure 17. Pulse techniques must be used that maintain the junction temperatures as close to the ambient temperature as is possible.

### Operating Characteristics, $V_{CC} = \pm 15$ V, $T_A = 25$°C (unless otherwise noted)

<table>
<thead>
<tr>
<th>PARAMETER</th>
<th>TEST CONDITIONS</th>
<th>MIN</th>
<th>TYP</th>
<th>MAX</th>
<th>UNIT</th>
</tr>
</thead>
<tbody>
<tr>
<td>$SR$ Slew rate at unity gain</td>
<td>$V_1 = 10 , V$, $R_L = 2 , k\Omega$, $C_L = 100 , pF$, See Figure 1</td>
<td>8°</td>
<td>13</td>
<td></td>
<td>V/µs</td>
</tr>
<tr>
<td>$t_r$ Rise time</td>
<td>$V_1 = 20 , mV$, $R_L = 2 , k\Omega$, $C_L = 100 , pF$, See Figure 1</td>
<td>0.05</td>
<td>µs</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$V_{IN}$ Equivalent input noise voltage</td>
<td>$R_S = 20 , \Omega$, $f = 1 , kHz$</td>
<td>18</td>
<td></td>
<td></td>
<td>mV/√Hz</td>
</tr>
<tr>
<td>$I_{IN}$ Equivalent input noise current</td>
<td>$R_S = 20 , \Omega$, $f = 1 , kHz$</td>
<td>4</td>
<td>µV</td>
<td></td>
<td></td>
</tr>
<tr>
<td>THD Total harmonic distortion</td>
<td>$V_{rms} = 6 , V$, $f = 1 , kHz$, $A_{VD} = 1$, $R_S \leq 1 , k\Omega$, $R_L \geq 2 , k\Omega$</td>
<td>0.003%</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

On products compliant to MIL-PRF-38535, this parameter is not production tested.
2.6. SECOND ORDER PHASE-LOCK LOOPS

LARGE-SIGNAL
DIFFERENTIAL VOLTAGE AMPLIFICATION
vs
FREQUENCY

EQUIVALENT INPUT NOISE VOLTAGE
vs
FREQUENCY