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AUTHOR   Andresen, S. G.
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ABSTRACT
                Presented is a set of laboratory experiments
developed to provide students with demonstrations and hands-on
experiences with a variety of basic communications methods. These
experiments may be used with students who have training in
engineering, as well as those with social sciences who have no
engineering background. Detailed exercises dealing with the
oscilloscope and the spectrum analyzer are provided for those
students lacking a technical background. It is assumed that the
laboratory will be taken with a theory course in basic electrical
engineering, simple mathematics, and communication theory. Each
experiment requires a lecture and a short demonstration preceding the
exercise, as well as considerable help and advice during the
laboratory period. (Author/SK)

BASIC EXPERIMENTS

IN

TELECOMMUNICATIONS

BY

S. G. ANDRESEN

ASSOCIATE PROFESSOR

ELECTRICAL ENGINEERING DEPARTMENT

UNIVERSITY OF COLORADO

JAN 1 9 1981

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PREFACE

The following basic experiments were developed with the help of NSF Grant No. HES74-20752 to go with the Master's Degree program in Telecommunications at the University of Colorado. The objective of this laboratory is to provide students with demonstrations and hands-on experience with a variety of basic communications methods. Our Master's program in Telecommunications involves work both in engineering and the social sciences and is designed to include students both with and without engineering backgrounds, and to provide a bridge across the social-technical interface. Since most of the students taking this laboratory lack a technical background, considerable time is spent on the oscilloscope and the spectrum analyzer. Those students familiar with the oscilloscope are allowed to proceed at a faster rate. All students take theory courses in parallel providing backgrounds in very basic electrical engineering, simple mathematics, and communication theory. The material is designed for a one semester course with most experiments requiring more than one meeting, and covers, as a minimum, the first 9 experiments, with the remaining experiments available for students progressing at a faster rate. Experiment 11 (Reception of cloud-cover pictures from weather satellites) provides an example of a simple complete communication system. Current information on weather satellites may be obtained by writing to the United States Department of Commerce, NOAA, National Environmental Satellite Service, Washington, D.C. 20233.

We have found that a suitable program consists of a relatively complete lecture and a short demonstration preceeding each experiment as well as considerable help and advice during the laboratory period. Many students also come in during non-scheduled times to repeat all or part of an experiment to get a better feel for the material and equipment. This we have encouraged. The reader will also notice that the material is more in the form of self-demonstrations than data-taking experiments. This we feel is better suited to the type of students expected in this program.

I would like to express my thanks to Dr. Carl Johnk for designing and writing Experiment 10 as well as all my colleagues for their helpful suggestions.

Boulder, Colorado, August 1976

S. G. Andresen

EXPERIMENT NO. 1

THE CATHODE-RAY OSCILLOSCOPE

Objective: To study the operation and measurement characteristics of the cathode-
ray oscilloscope.

Brief discussion of the oscilloscope. The cathode-ray oscillograph is a most
versatile tool and has tremendous importance in many branches of science and
engineering. Basically, it is a voltage-measuring tool, but when correctly
applied it will display waveforms of voltage or current, it will measure frequency,
time, phase angle, and it may even display the relation between two variables
directly on its screen and so it avoids having to perform a static point-by-point
plot. With appropraite transducers it will display sound waveforms, instantaneous
pressures, strains in materials as a function of time, and so on. The chief advan-
tages of the CRO are its high input impedance (it extracts very little energy from
the system being measured), its high sensitivity (achieved through the use of high
gain amplifiers), and its wide frequency-response range (it is sensitive to very
high oscillation freqi. cies of the input signal). The degree to which each of
these factors is exploited in a given CRO instrument depends upon the range of
uses for which that CRO was designed, of course.

Generally speaking, a cathode-ray oscilloscope (abbreviated "CRO" in the
following) is an electronic instrument that can be used for visually displaying
a voltage function v(t) as a graphical function of time. The visual image on
the cathode-ray tube (abbreviated "CRT" in the following) is obtained as a result
of the bombardment of a fluorescent mineral coating on the inside of an evacuated
glass envelope by means of a low-inertia electron beam, which has been focused by
means of electrostatic lenses into a small spot by the time the beam reaches the
flourescent coating. A necessary property of the coating is that it shall emit
visible light rays when so bombarded. Suitably synchronized deflections of the
beam spot in the x- and y-directions on the CRT face can then be brought into play
such that the beam spot will describe on the CRT face a path which is a replica of

Horizontal Deflection
Voltage (assisted by
amplifier)

v_H

Cathode

P_1

P_4

P_3

P_2

v_V

Beam Spot position when
v_V is positive and $v_H=0$.

Beam Spot position
for $v_H=v_V=0$

Vertical Deflection Voltage
(assisted by amplifier)

y

x

Principal axis
of CRT

CRT face
(inside coated)

Beam Spot position when
v_H and v_V are positive

Beam Spot position when
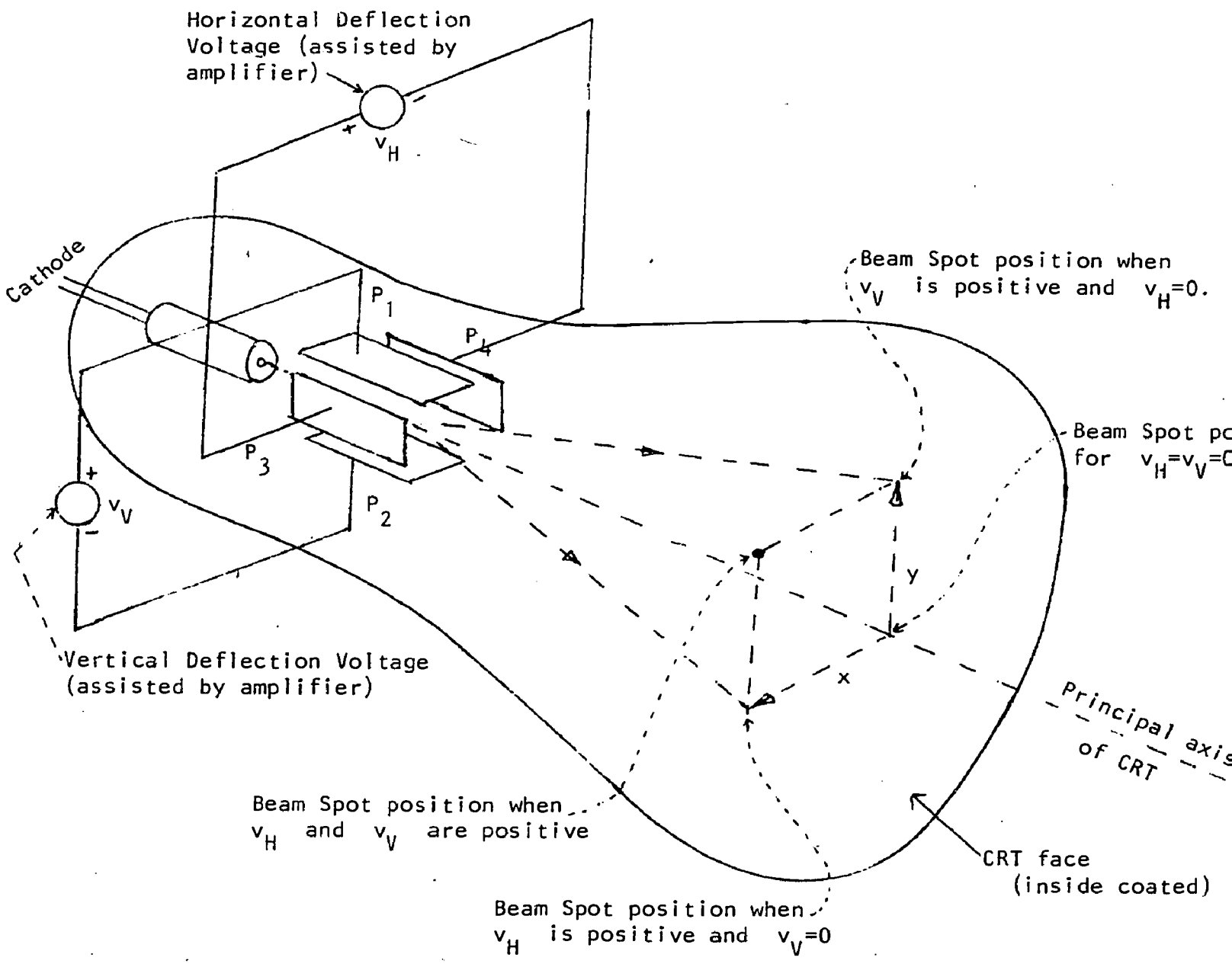v_H is positive and $v_V=0$

<u>Fig. 1</u> Diagrammatic representation of a cathode-ray tube, showing electron gun
or source of a high-energy electron beam, deflecting plate-pairs whose
electric fields are disposed at right angles, and effect of different
potential differences v_H and v_V upon beam spot position on CRT face.

Deflections x and y are proportional to voltage differences v_H and
v_V in a well-designed tube.

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some voltage function v(t) . The beam spot deflections can be introduced at
straight angles to the electron beam axis by means of the electric-field forces
produced by pairs of metal parallel-plates, as represented in Figure 1. A know-
ledge of electron ballistics and the field forces present is essential to a complete
study of this device.

Some CRT's have more than one electron gun and so can produce more than one
beam on the CRT face at a given instant of time. A two-gun CRT is known also as
a "dual-beam CRT," for example. Tubes having as many as eight guns have been built
commercially.

Every CRT has a property known as "deflection sensitivity," which is a measure
of the voltage v_H or v_V required to deflect the spot a certain distance, say
one centimeter, from the center position. A typical deflection sensitivity may
be 30 volts per centimeter, so to deflect the beam spot 8 centimeters from the
center position on the 5-inch diameter face of such a CRT would require 240 volts,
which is a relatively large deflection voltage. If this CRT were to be used to
measure a 10-millivolt signal in a circuit, a voltage amplifier having a voltage
gain of 24,000 (almost 88 db) would be needed in order to secure an 8-centimeter
spot deflection on the CRT. In fact, calibrated voltage amplifiers are used in
practice, such that by switching from one voltage gain setting to another, different
deflection sensitivities are made available to enable the CRO instrument to measure
voltages over a wide range of values. This arrangement permits this CRO to be used
as a direct-reading, calibrated voltmeter (i.e., a meter, or measurer, of voltage)
having high sensitivity and a very wide range.

At this point let us consider how one would measure a voltage of the form
v(t) = 10 sin ωt . If the angular frequency, ω , is very small, this is a slowly
varying voltage and may be read by a fast observer with a d.c. voltmeter. The
result could be plotted as v(t) vs. t . As the frequency is increased, the
voltmeter becomes inadequate. Then a recorder which moves paper under a moving
pen could be used. This would give a plot of v(t) vs. t directly. But as the
frequency is increased still higher, the recorder becomes inadequate. The CRO,
which uses a low-mass electron beam, now becomes very suitable. Again it is desir-
able to have a plot of v(t) vs. t directly, if possible.

The v(t) versus t display is usually presented on the CRT face in rectan-
gular-coordinate form, implying that the x-axis, or horizontal axis, represents a

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linear time scale on the face of the CRT, while the y-axis, or vertical axis, repre-
sents the linear v(t) scale on the face of the CRT.

Let us consider methods for deflecting the beam horizontally as a linear
function of time. Such a function would be $v(t) = kt$. However, this voltage
would eventually deflect the beam off the right side of the scope. Then some
method must be found for returning the beam to the starting point and sweeping
again. A triangular wave is such a function. This would sweep the beam from the
left side across to the right side at a constant velocity and then return to the
left side at the same velocity. But this is not satisfactory as the return portion
will be displayed over the first half and it may not be easy to tell which is which.

The sawtooth wave shown in Fig. 2a has a linear increase to its maximum value
and then a rapid return to its starting value. The beam returns to the left side
much faster than it progressed toward the right side of the screen and thus the
return trace is less bright than the forward trace. However, it is better to elec-
tronically extinguish or "blank" the beam during its return path and this is done
in modern CRO's.

(a) Synchronized sweep with sawtooth wave.
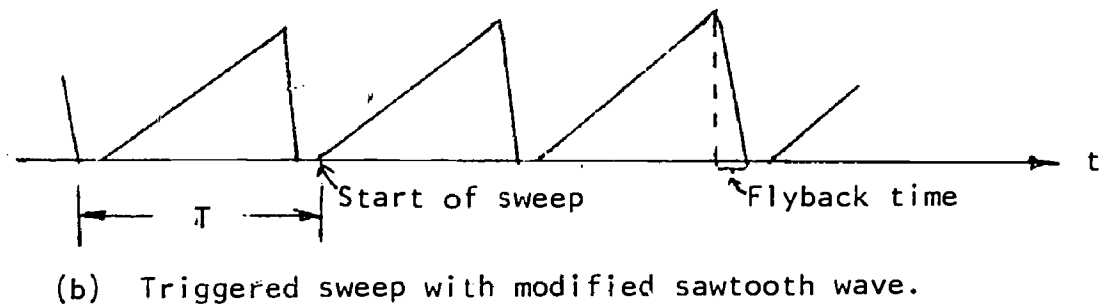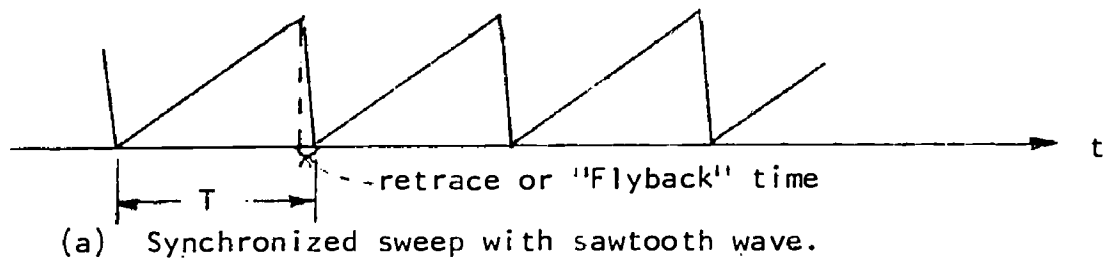
(b) Triggered sweep with modified sawtooth wave.

Fig. 2. Voltage waves used for horizontal deflection or
"sweep" of electron beam.

The period, T, of the wave of Fig. 2a must be equal to some integral number
times the period of the voltage wave applied to the vertical deflection plates.

Otherwise, the beam would not retrace the same path on successive traverses and the screen would not show a clear single curve. This means manually adjusting the period of the sweep circuit until synchronization is obtained. A more convenient method of obtaining a clear single trace is by using a modified sawtooth wave shown in Fig. 2b, in which the start of the sweep is "triggered" by a specific value, and slope, of the voltage applied to the vertical deflection plates.

Next we need a method of deflecting the beam in the vertical direction proportional to the input function $v(t)$. This is done simply by amplifying the function $v(t)$ a calibrated amount and applying the voltage to the vertical deflection plates. The voltage value needed to provide a certain y-deflection will depend on the characteristics of the CRT and of the vertical amplifier, and this calibration, usually given in volts per centimeter of y-axis (volts/cm.), can for our purposes be read off the voltage-gain knob, or sensitivity knob, of the vertical amplifier in question. Now we can use Fig. 3 to demonstrate how a wave is displayed on the scope face. Point 0 corresponds to $t = 0$ for both $v_s(t)$ and $v(t)$. Then at $t = t_1$, the sawtooth has moved the beam to the right and $v(t)$ has moved the beam up 2.5 cm. This gives Point 1 on the scope. And so forth on out to $t = t_2$ which corresponds to Point 2 on the scope face.

The beam returns to its starting point at $t = t_3$ and stays there until the voltage, $v(t)$, has a certain magnitude and slope. In Fig. 3 this occurs when $v(t) = 0$ and has a positive slope.

The student is not expected here to understand fully the reasons why the panel controls of the CRO do as they do, but he is expected to be able to use the controls properly so as to produce the desired effect on the CRT.

Refer to the instruction manual for the characteristics and operating information of Tektronix type 543 oscilloscope with type CA plug-in unit.

Suggested Procedure:

1. Turn on the Tektronix 543 oscilloscope. Several minutes of warmup may be required to stabilize the d-c amplifiers. CAUTION: When the beam spot is stationary or moving at very slow speeds, keep the INTENSITY to a low value to avoid burning the CRT screen. Focus the spots or trace appropriately.
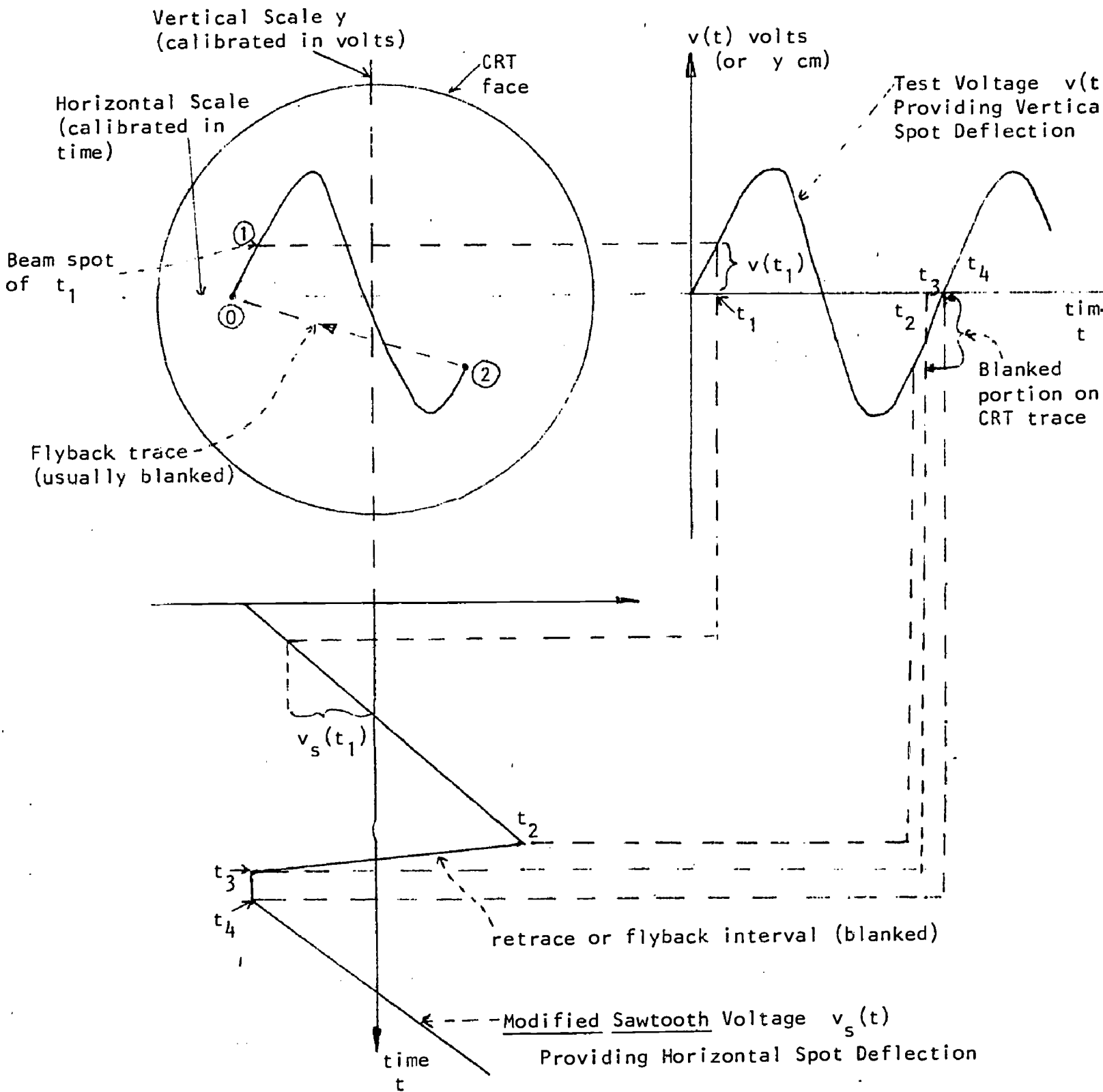
Fig. 3 Showing how a modified sawtooth voltage $v_s(t)$ which provides horizontal spot deflection, and a test voltage $v(t)$ (waveform to be observed) which provides vertical spot deflection, are combined to produce the $v(t)$ versus t display on the CRT face. Spot deflections or displacements on the CRT face are actually only _proportional_ to $v_s(t)$ and $v(t)$; this is a property of the linear voltage amplifiers used in the CRO.

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Observe the beam spot motion for a number of different time-base rates, for no input signal applied. Note that the TRIGGERING LEVEL control must be on AUTOMATIC or the STABILITY all the way clockwise to obtain a CRT trace, if no signal is applied. Note that one may use a watch second-hand to check roughly the calibration of the slowest sweeps. Try this.

2. <u>Observation of voltage waveforms</u>. Connect the CH-748 function generator to one of the inputs. Set the Triggering Level control knob at Automatic and switch the Sensitivity control to an appropriate position compatible with the expected output voltage from the generator. Since the CRO will measure voltage from input to the chassis (or ground) of the CRO, it is important that the chassis (or ground) of the function generator be connected to the chassis of the CRO. Set the frequency of the function generator at 1 kHz and display a triangular wave on the CRT. Adjust the Sensitivity control of the CRO and the Amplitude control of the function generator so that the wave has a 6 centimeter peak-to-peak deflection with the waveform centered on the screen. Adjust the sweep rate so that only one to two cycles of the wave are shown. By switching the function generator, observe the sinusoidal and square waves, also.

3. <u>Check of waveform frequency against calibrated time base</u>. With the generator frequency set at 100 Hz for a triangular wave, use the calibrated time base to check this frequency. Assume the CRO to be the "standard." Make this check for two ranges of the generator at this same frequency, (that is, 100 x 1 and 10 x 10), and calculate the percentage deviation from the standard in both cases. Repeat this check, one range only, at 10 Hz and at 1 kHz.

4. <u>Use the square-wave Calibrator</u> to check the calibration of the Sensitivity control. Make a check at all settings of the calibrator.

5. <u>Use of Sweep Magnifier of horizontal display</u>. For a suitable triangular wave displayed on the CRT, observe the effect of the Sweep Magnifier on the display. Compare this effect with that produced by using a faster Time Base rate and comment.

6. <u>Observation of effect of Trigger Level Control</u>. With a triangular voltage wave ovserve the effect of moving the Trigger Level Control from the Automatic

position. By means of a series of sketches describe what you observe. Is the effect the same whether the Trigger Selector control knob is set at a.c. or d.c.?

With the input selector knob set at d.c., observe the effect of varying the Trigger Level Control on the start of the sweep of this waveform. How does a change in d.c. value affect the start of the sweep? Are these effects the same if the Trigger Selective control knob is shifted from the d.c. mode to the a.d. mode?

8. Use of the X-Y plotter arrangement. To illustrate this feature inject the same signal into one of the inputs and into the External Horizontal Input. Set the Horizontal Display to Ext.

Note that at low frequencies, with input selector knob at a.c., the trace will not be what you expect. This is because of the effect of the capacitor in series with the input for this connection.

If time permits, use two separate generators to obtain X-Y plots. Try sinusoidal, triangular, and square waves. You may need to join with another squad in doing this portion of the experiment.

Also if time permits, connect a series RC circuit of the function generator. With a sinusoidal input show the ellipsoidal X-Y plot obtained if one voltage is the applied voltage and the second voltage is across either R or C. Observe the change in the ellipse as frequency is varied.

9. DC-Balance. The purpose of the DC Balance is to permit changes in the Sensitivity Control without changing the vertical position of the trace. The student should become familiar with the technique for setting the DC balance.

With an input probe connected to chassis, and with the Sensitivity Control knob set at a low level of sensitivity, adjust the Position control knob so that the trace is at a convenient vertical position. Next, increase sensitivity and if the trace changes, then adjust the DC Balance so that the trace is returned to its original position. Continue until maximum sensitivity is reached. You should now observe that the trace stays at the same vertical position as sensitivity is varied.

EXPERIMENT NO. 2

THE SPECTRUM ANALYZER

Brief discussion of the spectrum analyzer

A spectrum analyzer is an instrument that graphically presents a plot of amplitude as a function of frequency for a selected portion of the spectrum.

One could in principle build a spectrum analyzer using a band-pass filter (BPF) whose center frequency could be varied and plotting the filter output amplitude vs. its center frequency.
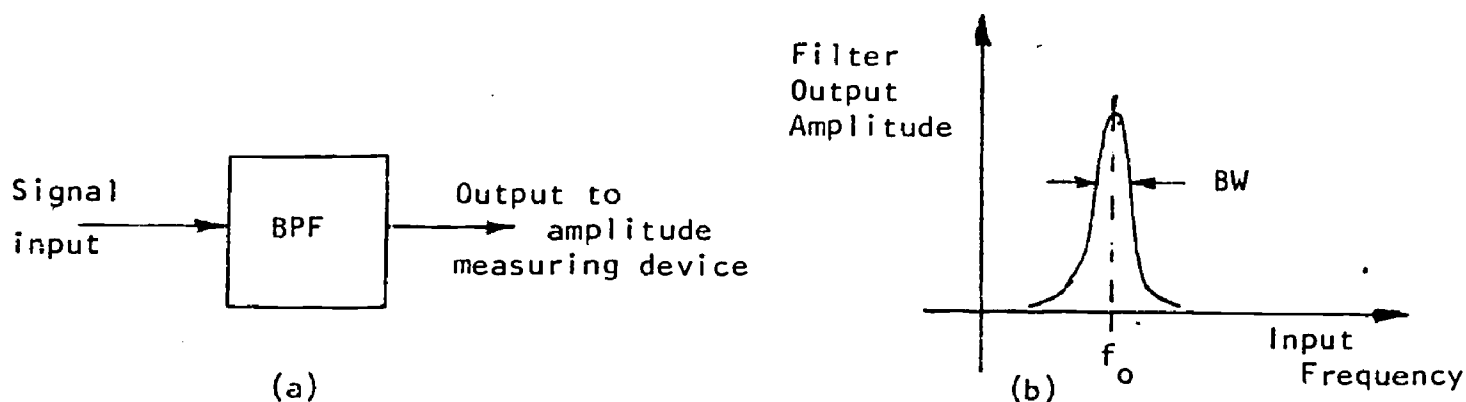
Fig. 1. BPF and its characteristic, f_o is the filter center frequency. The filter output amplitude is plotted vs. the frequency of a constant amplitude sinewave input signal.

Fig. 1 shows the BPF and its characteristic which is obtained by slowly varying the frequency, while holding the amplitude constant, o an input sinewave and plotting the output amplitude vs. the varying frequency. Note that the same shape plot would have been obtained if the filter center frequency, f_o, were varied, while impressing a fixed frequency sinewave at the filter input and plotting the output amplitude vs. the filter center frequency. In this case

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the center of the response curve shown in Fig. 1(b) would coincide with the input sinewave frequency.

The bandwidth (BW) of the BPF has to do with the resolving power or resolution of this "spectrum analyzer." This may be seen by applying two sinewaves of different frequencies and amplitudes simultaneously to the filter and plotting the output vs. f_o . This is done in Fig. 2 for two different BW'
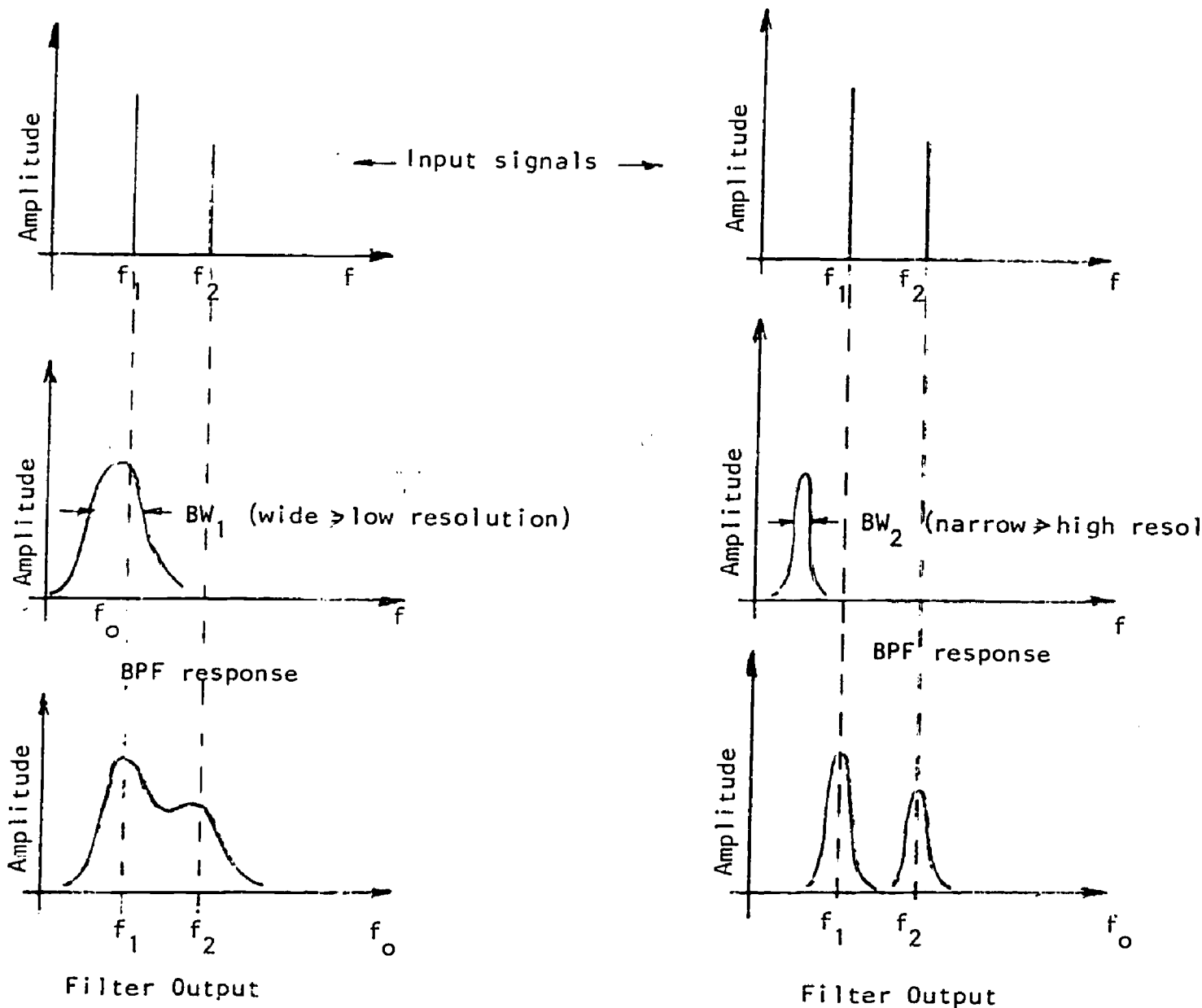

Fig. 2. Effect of filter BW on resolution

A variable center frequency BPF is implemented by combining a voltage controlled oscillator (VCO), a mixer (or converter), and an intermediate frequency amplifier (IF amplifier). A block diagram is shown in Fig. 3.
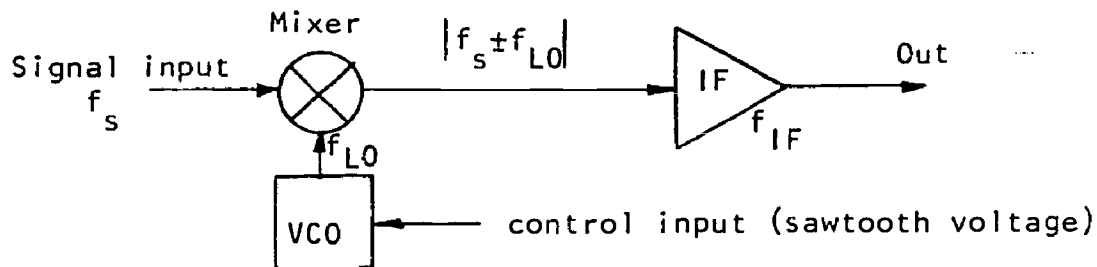
Fig. 3.   Implementation of a variable center frequency BPF.
           With a sawtooth control input the VCO frequency,
           f_{LO} , will be swept.

The VCO is often called the local oscillator (LO). With a signal input at f_s Hz and a local oscillator at f_{LO} Hz, the output of the mixer consists of two signals, one at the sum frequency $f_s + f_{LO}$, and the other at the difference frequency $|f_s - f_{LO}|$. The IF amplifier is a fixed center frequency BPF with gain. In the IL5 plug-in spectrum analyzer the IF center frequency is 3 MHz, and the VCO is swept from 3 to 2 MHz by the horizontal sweep voltage from the 543 oscilloscope main frame. The IL5 is designed to provide a spectral display within the range 50 Hz to 1 MHz, thus the IF amplifier selects the <u>sum</u> frequency of the input signal and the local oscillator, and rejects the difference frequency. Consider now what happens if the input signal is a sinewave of frequency f_s = 0.5 MHz. The LO sweeps from 3 MHz (startpoint) to 2 MHz (stoppoint). When the LO frequency has reached f_{LO} = 2.5 MHz the sum $f_s + f_{LO}$ = 3 MHz = f_{IF} and there appear an output at the IF. For any $f_{LO} \neq 2.5$ MHz, there is essentially zero (or very little) output of the IF. Thus, the system in Fig. 3 is equivalent to a swept center frequency BPF with a center frequency range of 0 (DC) to 1 MHz. This system is the same as used in your AM pocket transistor radio with the difference that in the radio the LO is tuned manually for station selection.
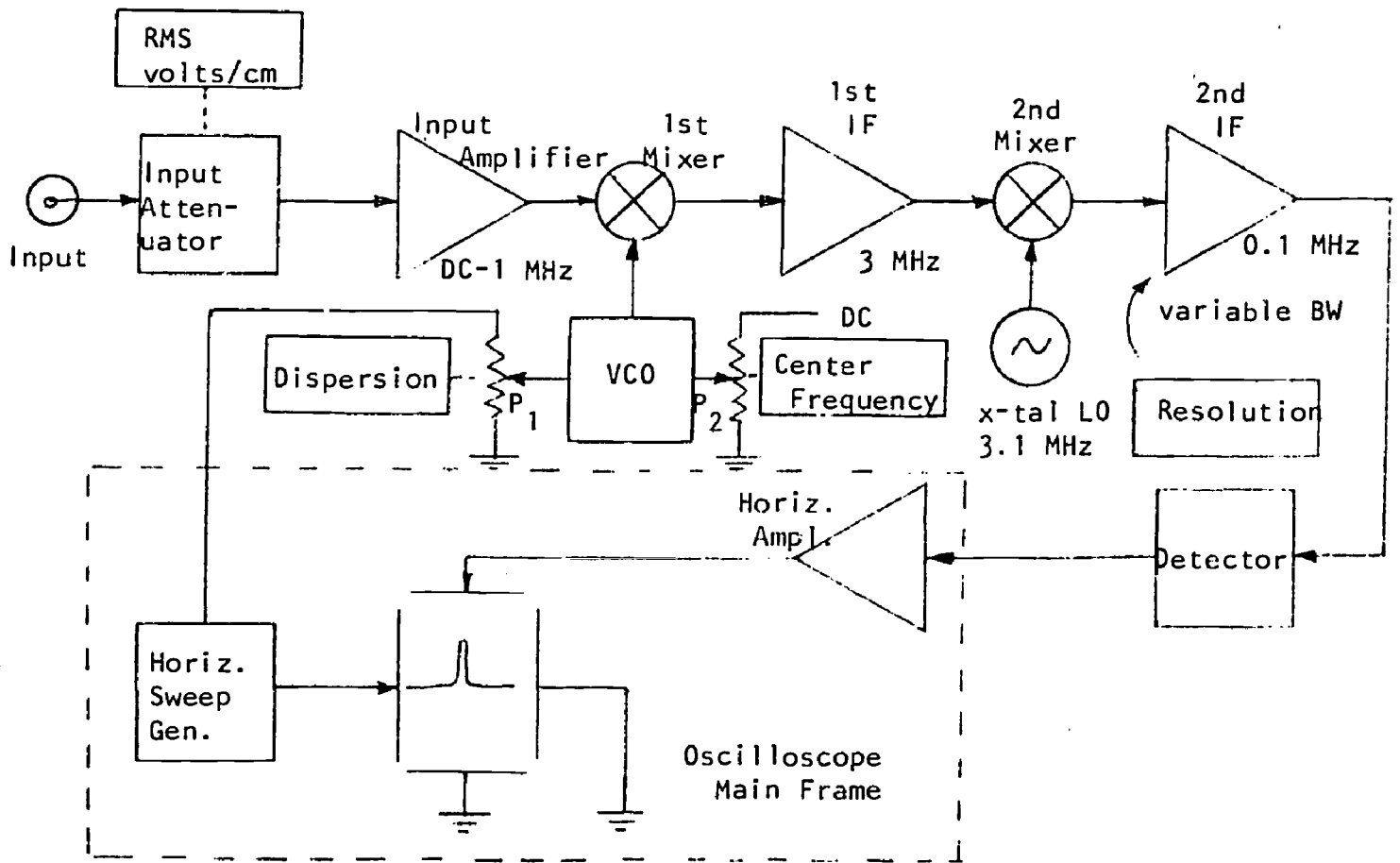
Fig. 4. Simplified block diagram of IL5 spectrum analyzer.

In order to get variable BW, or variable resolution, a 2nd mixer, LO, and IF is used. The 2nd IF has variable BW. Following the second IF is a detector whose output drives the oscilloscope vertical amplifier. Thus, the y-axis of the display corresponds to amplitude. Since the sweep voltage drives both the VCO and the horizontal amplifier, the x-axis of the scope display corresponds to frequency.

A simplified block diagram of the IL5 spectrum analyzer is shown in Fig. 4.

The maximum sweep range of the VCO is from 3 to 2 MHz. This range may be reduced by the front panel control labeled DISPERSION (P_1 in Fig. 4). The control labeled CENTER FREQUENCY (P_2 in Fig. 4) refers to the frequency displayed in the center of the CRT face (or in the approximate middle of the horizontal sweep). Thus, with these two controls one can select any portion of the maximum range for display. A few examples will illustrate this:

Center Freq.	Dispersion	Display Range (10 cm horiz. sweep)
500 kHz	100 kHz/cm	DC 500 kHz 1 MHz
500 kHz	10 kHz/cm	450 kHz 500 kHz 550 kHz
50 kHz	10 kHz/cm	DC 50 kHz 100 kHz

Notice that when 0 Hz (DC) is included in the range the VCO starts sweeping at 3 MHz. This start frequency is the same as the 1st frequency and since the mixer is not perfect, some of this 3 MHz signal will leak through and be amplified and will show up at the beginning of the sweep as a small pip. This pip is very useful since it indicates the zero Hz frequency point of the display and is therefore called the 0 Hz marker.

Procedures

1. The IL5 has several front panel screwdriver adjustments so that it can be adjusted and trimmed correctly. For first time use and for learning how to use it, it is useful to go through the adjustment procedure. Accordingly, follow the instructions and procedures on pp. 2-1 through 2-15 of the IL5 manual.

2. Display the spectrum of a sinewave for various frequencies and amplitudes. Check the analyzer calibration in frequency and rms volts/cm. Assume that your generator is the "standard." Notice what happens if the sweep speed is too high.

3. Display the spectral components of a triangular and a rectangular wave. Repeat for a pulse train and notice the $|\sin x/x|$ envelope of the spectrum (the spacing between the spectral lines should be 1/T, where T is the period of the pulse train, and the first zero crossing of the envelope should occur at $1/\tau$, where τ is the pulse duration). Be sure to use a sufficiently slow sweep speed. (If you are using two scopes you can observe the signals in both the time and frequency domains.)

EXPERIMENT NO. 3

BASEBAND SIGNALS AND SIMPLE LOW-PASS CHANNELS

Reference: A. Bruce Carlson: "Communication Systems," 2nd Ed. McGraw-Hill, 1975.

Introduction

The direct electrical output of a transducer is usually called a baseband signal. An example is the output of a microphone or a strain gauge. Baseband signals are usually in the audio frequency range (not always true -- the high speed output of a computer can also be considered a baseband signal). Baseband signals are of two types: analog and digital; with the analog signals occurring more often since most transducers are analog in nature. You have already displayed a number of simple periodic baseband signals (sine, triangular, and square waves out of a function generator, and pulse trains out of a pulse generator) on a scope. In the following we will also consider speech, music, and digital data signals. Baseband signals may be transmitted directly over a baseband channel without processing (such as modulation) at least from the transducer to a possible processor. A question then arises as to how wide (in frequency) a channel is necessary for the direct transmission of these baseband signals, say for a given amount of distortion? This baseband bandwidth (BW) requirement is also important in that it determines the final BW of the processed baseband signal.

Procedures

a) <u>Amplitude</u> <u>and</u> <u>phase</u> <u>distortion.</u> (See pp. 139-153 in reference.)

Generate a 1kHz square (nearly) wave using the CH-748 function generator. Such a square wave consists of a fundamental frequency of 1kHz plus odd multiples (harmonics) of 1kHz (at 3, 5, 7, ... kHz). There are no Fourier components below 1kHz, assuming this to be a symmetrical square wave (no d.c. component). Using a variable Krohn-Hite filter set in the high-pass (HP) mode with the low frequency cut-off (3dB point) at 0.5kHz, look at the filter output. Is it a "good" square wave? Check the filter performance by switching to a 1kHz

sine wave and simultaneously observing the filter input and output. Trigger the scope externally from the function generator. Is there any amplitude attenuation at this frequency? How about phase shift? Can you now explain the square wave distortion you got?

If you were listening to the two square waves (filter input and output) over a speaker or earphones would they sound different? Try it.

Now set the filter in its low-pass (LP) mode and vary the high-frequency cut off. Approximately how wide (BW) must this low-pass "channel" be in order to approximately preserve the square wave shape?

b) <u>HP and LP filter speech.</u>

Note that the intelligence (information) resides mostly in the high frequency components, while the volume is in the low frequency components. Determine the minimum LP channel BW necessary for reasonable sounding speech. (Telephone-quality voice has a BW of about 200 to 3200 Hz with a signal-to-noise ratio of 25-35 dB.) Repeat for music.

c) <u>Signaling speed vs. BW.</u> (See pp. 158-161 in reference.)

A digital signal is often thought of as consisting of a string of discrete-amplitude pulses. This is often the way it comes from the source (for example, the output of a computer). Fig. 1 shows a few possible idealized digital wave forms.
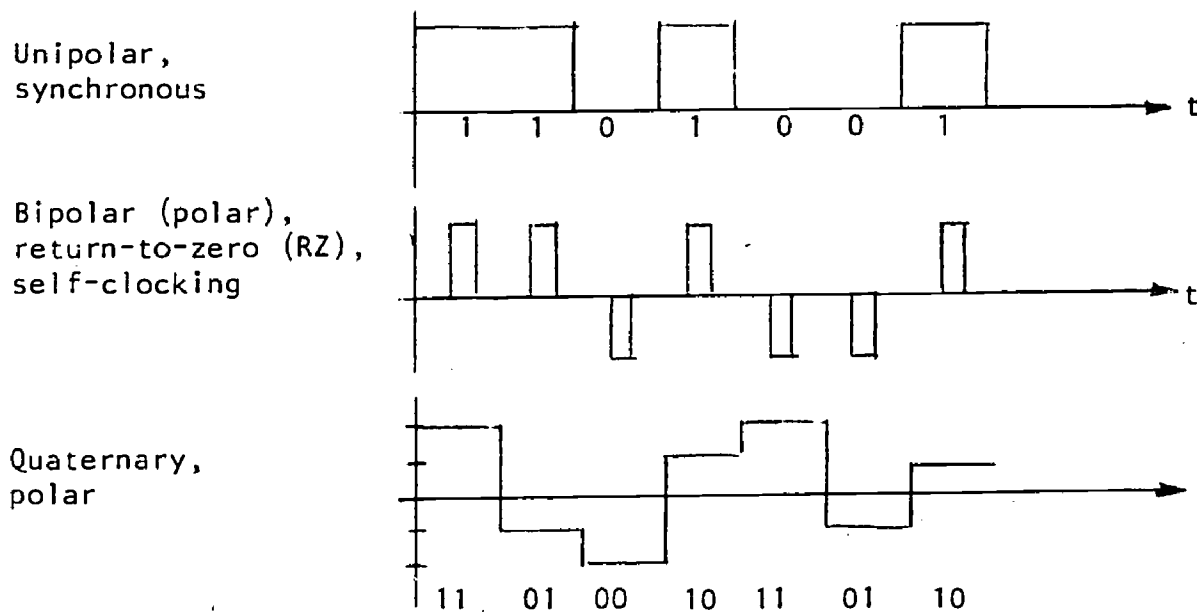
Fig. 1. Some idealized digital wave forms.

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if the pulse-to-pulse spacing is τ sec, then

$$s = \frac{1}{\tau} \text{ symbols/sec}$$

is called the signaling speed (or rate). The information rate (bits/sec) is equal to s if we use a binary (2 level) signal, but is larger than s for a multilevel signal. With a channel of finite BW s is limited, and according to Nyquist:

Given an ideal low-pass channel of bandwidth BW, it is possible to send independent symbols at a rate $s \leqslant 2BW$ symbols/sec without intersymbol interference. It is not possible to send independent symbols at $s > 2BW$.

In practice the rate may be considerably less than the maximum (2BW) due to channel imperfections.

In order to test a channel we may use a display technique called an eye-pattern. This eyepattern will give us an indication if a threshold (or decision) device connected to the channel output will function properly, i.e. is capable of deciding (with low errors) at a given clock time if the channel output is, say, a "zero" or a "one" in the case of a binary signal.

An eyepattern is set up by triggering the scope at the clock rate and adjusting the sweep to last one time slot. Fig. 2 shows various patterns obtained.

CRO
pattern:
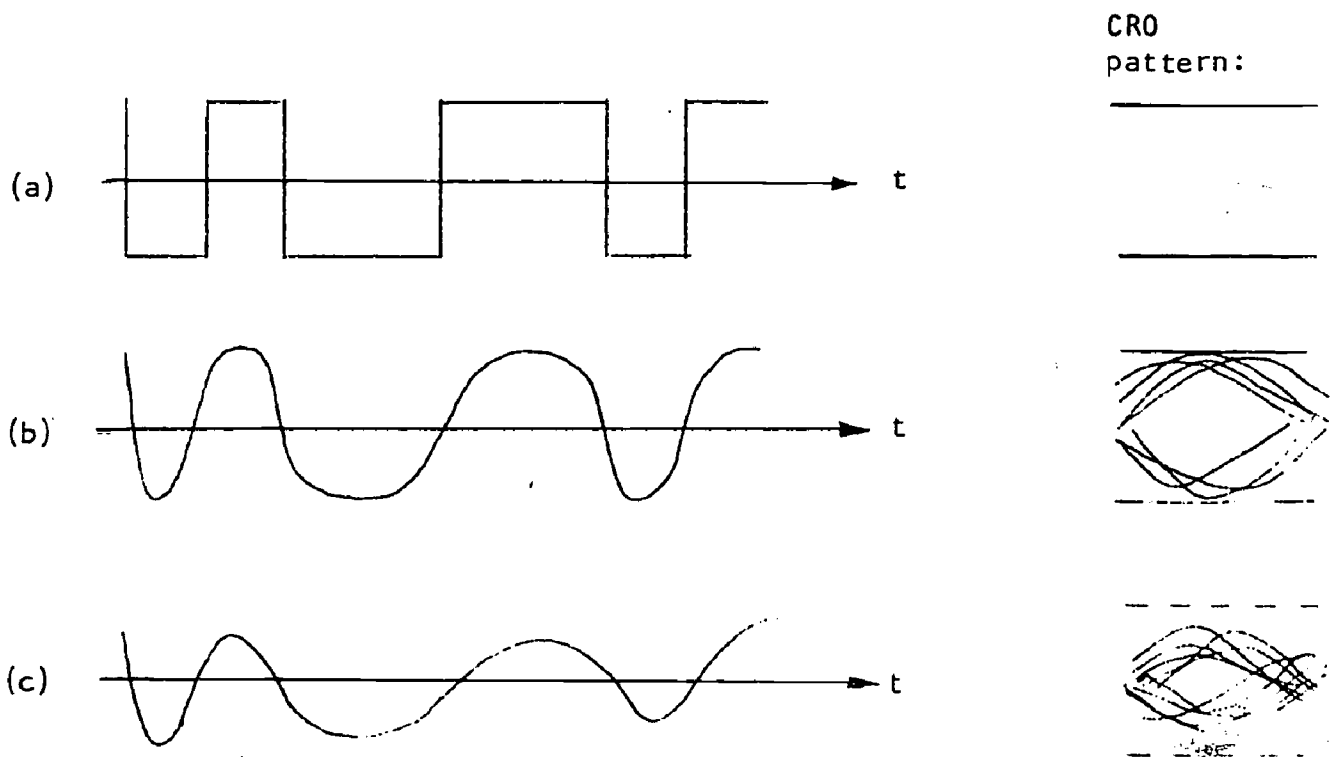
Fig. 2. Bit-trains and corresponding eye patterns.

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In Fig. 2(a) there is no BW limitation and the bit-train waveform would appear as on the left and the scope pattern as on the right. Fig. 2(b) illustrates the effect of limiting the BW. In (c) further BW limiting is indicated, and the open space at the center of the "eye" has almost closed. In the case of an almost closed "eye" a threshold device would make a large number of errors. A multilevel signal would give rise to several "eyes."

In order to get a realistic eyepattern it is necessary to use a random signal. To demonstrate, set up an eyepattern display using a 4-level pseudorandom sequence generated by a pseudorandom code generator (PRCG) as indicated in Fig. 3. Vary the LPF cut-off frequency and observe the eyepattern. What does the pattern look like at s = 2 BW ?
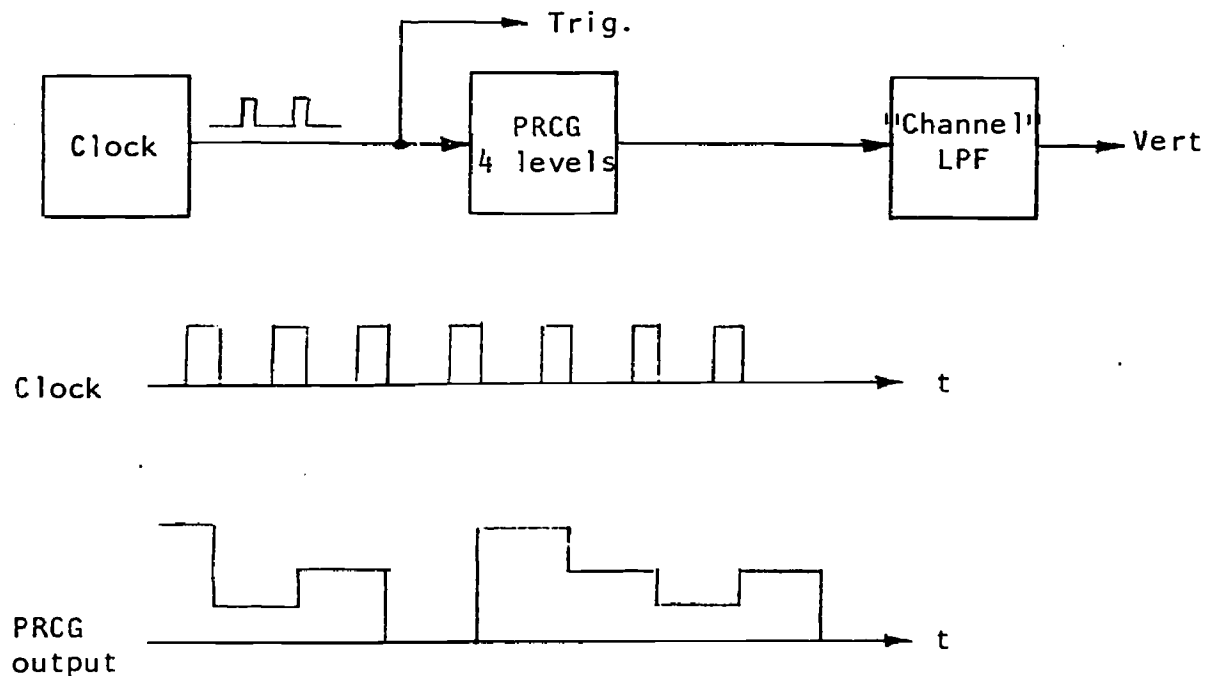
Fig. 3.

Distortions produced by an imperfect channel can, to a certain extent, be undone by a receiver filter called an equalizer. To adjust an equalizer it is often necessary to precede each message by a test bit stream and adjust the equalizer for maximum eye openings.

A simple channel with echo and its equalization will be demonstrated in class.

EXPERIMENT NO. 4

SAMPLING, QUANTIZATION, AND ENCODING OF BASEBAND SIGNALS

Reference: A. Bruce Carlson, Chapter 8

Introduction

Analog baseband signals are often sampled, quantized and encoded into a digital code before further processing (such as modulation) and transmission. Sometimes all of the above forms are done, other times just the sampling operation is performed. A basic requirement for these operations is that the baseband signal be bandlimited; that is, its bandwidth be limited from DC to some maximum frequency f_m . In this case the sampling theorem states

> Any 2 f_m independent samples per second will completely characterize a bandlimited signal; or, bandlimited (DC-f_m Hz) signals may be sampled at a rate equal to or greater than 2 f_m Hz and reconstructed from their samples without errors (in the absence of noise).

The minimum sampling rate (2 f_m) is called the Nyquist rate. One may also show that finite width natural sampling does not introduce any errors.

Fig. 1 shows the various time and frequency domain forms of a typical sampling operation (using natural sampling, i.e. the top of the samples correspond to the sampled waveform).

Considering the frequency domain of the samples of x(t) , one sees that ideal low-pass filtering (LPF) recovers the original spectrum, X(f) , and hence the original waveform, x(t) .

To avoid distortion, one must have that

$$f_m \leq f_s - f_m ,$$
(1)

otherwise the spectral components overlap and LPF will not yield the original spectrum.

(1) is equivalent to

$$f_s \geq 2 f_m$$
(2)

22

Time Domain: Frequency Domain:
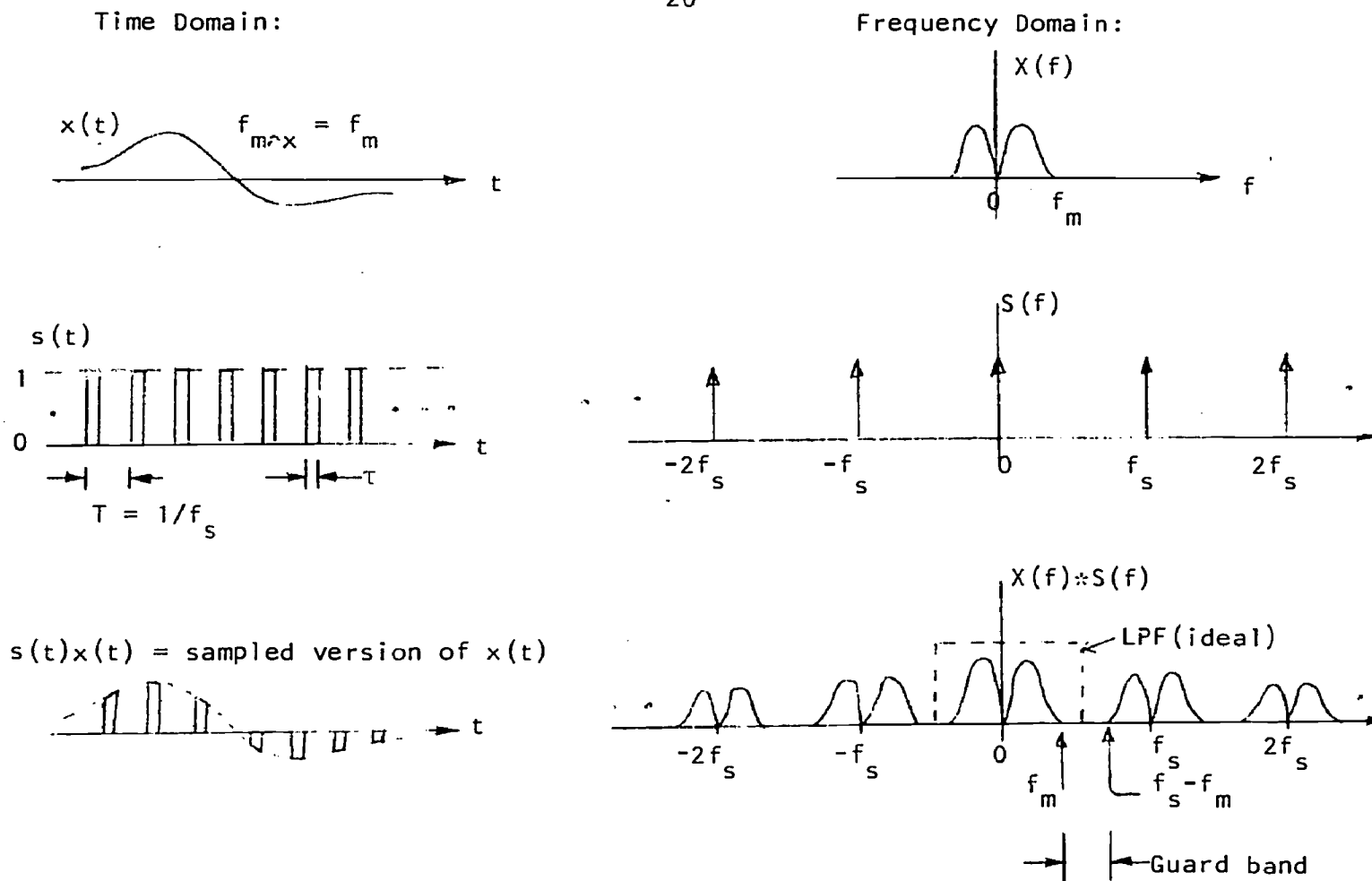
Fig. 1. Typical natural sampling

In practice one needs a guard band between the spectral components since it is not possible to build an ideal LPF as shown in Fig. 1, thus it is necessary to sample at a rate somewhat higher than the minimum ($2 f_m$).

Secondly, it is not possible to have a <u>completely</u> bandlimited signal as indicated in the figure; the spectrum will usually have a small tail beyond f_m, this again means a higher sampling rate than the minimum theoretical. (In the telephone system sampling is done at 8 kHz on voice signals limited to approximate 3.2 kHz, thus producing an approximate guard band of $8 - 6.4 = 1.6$ kHz.)

Notice that the pulse length, τ, of the samples does not affect the <u>shape</u> of the spectrum of the sampled version, hence does not introduce any errors in the reconstruction. Changing τ, however, will change the level (or "volume") of the reconstructed signal since increasing τ increases the energy per sample.

(If τ is too short the reconstructed signal may disappear in noise, when noise is present.)

Pulses of the shapes of Fig. 1 cannot be used in practice because of their high frequency content (fast rise and fall times and sharp corners); instead a certain basic pulse shape, p(t) , is used and the resulting sampling operation is called practical sampling, Fig. 2.

Basic pulse shape

Original signal

Spectrum of basic
pulse shape

Sampled version
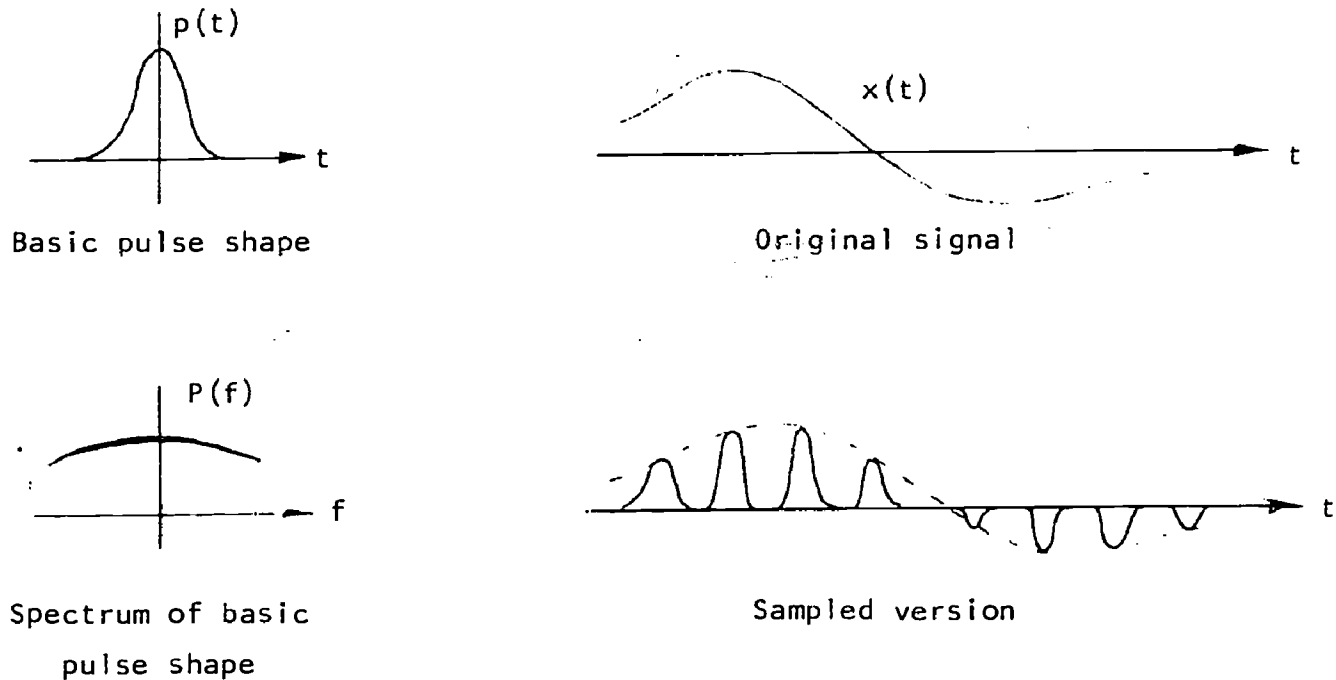
Fig. 2. Practical Sampling

The effect of using this pulse shape is like filtering of a naturally sampled signal with a filter having a transfer function P(f) which can be compensated for at the receiver by a receiver filter of transfer function

$$H(f) = \frac{1}{P(f)} \quad . \tag{3}$$

See reference for details.

Notice that for short samples there is a relatively long dead time on the channel. This dead time could be used for the transmission of a 2nd set of samples belonging to another bandlimited signal; in fact, not just an additional

signal may be sent, but several signals. Such a system, in which the samples of several signals are interleaved in time, is called time division multiplexing (TDM).

To implement sampling or TDM one needs an electronic commutator as indicated in Fig. 3.
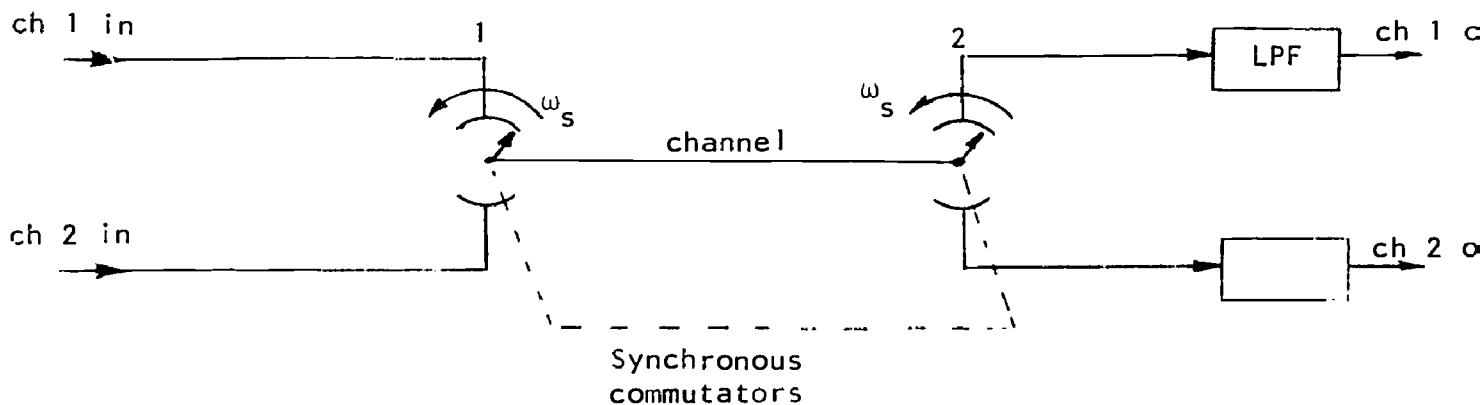
Fig. 3.   Sampling and TDM. Commutator 1 samples
each channel at a rate $f_s/2$ samples/sec,
while commutator 2 distributes the samples.

Procedures

(a) Set up a sampling and reconstruction system as in Fig. 4, and investigate its performance using an audio-frequency sinewave for $x(t)$. Next test the system using music for $x(t)$ and listen to the reconstructed waveform over a speaker. Check the effect of varying the sampling rate, f_s , and the sample widths, τ . Make sure you include $f_s = 2 f_m$ where f_m is the LPF cut-off frequency.
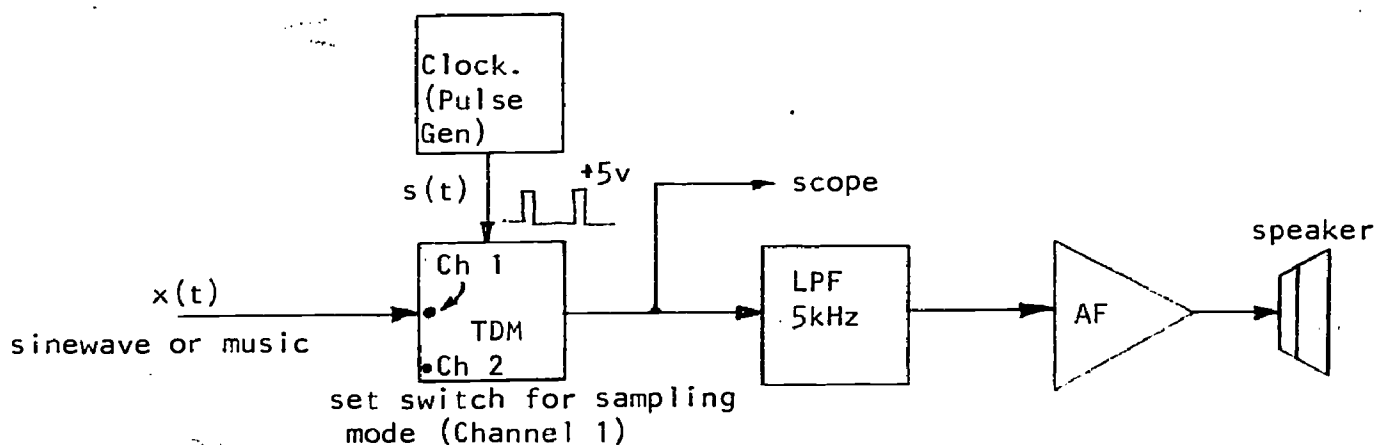
Fig. 4. Natural sampling of audio waveforms.

(b) Set up a 2 channel TDM with, say, a sinewave input for one channel and a triangular wave input for the other channel. Display the time domain form of the multiplexed signal and check the reconstruction for each channel. See Fig. 5.
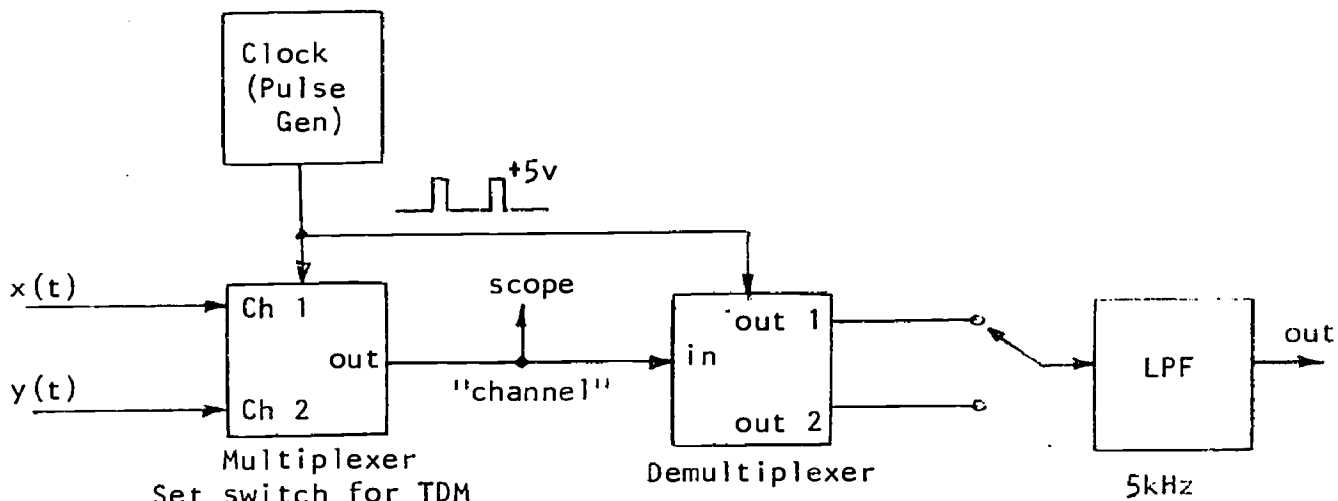
Fig. 5.   2 channel TDM/Demultiplexer system.

In practice the clock pulses are transmitted along with the multiplexed signal over the channel for synchronization information for the demultiplexer. Such a system is available with a fixed clock rate of 20 kHz (f_s = 10 kHz per channel). You may try this if you like. See Fig. 6.
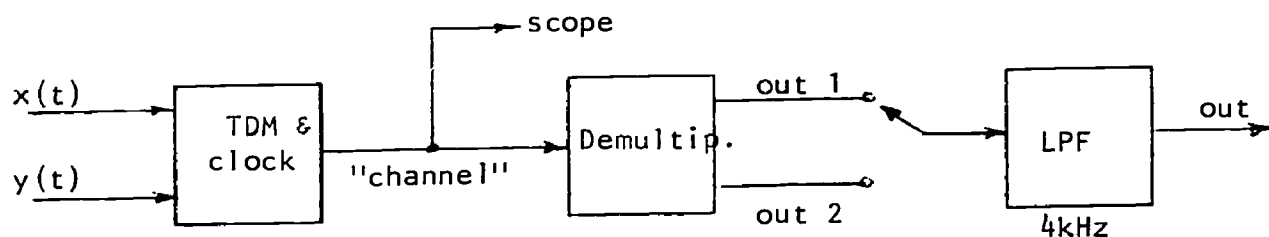
Fig. 6.   2 channel TDM with clock pulses transmitted along with the signals.

In sampling, the information about x(t) is carried by the amplitudes of the samples, and thus a sampling operation is sometimes referred to as pulse amplitude modulation (PAM). Since the pulse duration τ is constant then one may think of

the areas of the pulses as the information carriers. Instead of using constant width pulses one could use constant amplitude pulses and let the widths vary according to the sample values. This is referred to as pulse duration modulation (PDM). Again the information will be carried by the pulse areas, and reconstruction of $x(t)$ may therefore be done by a LPF as in the case of PAM.

Set up and test a PDM system as in Fig. 7.
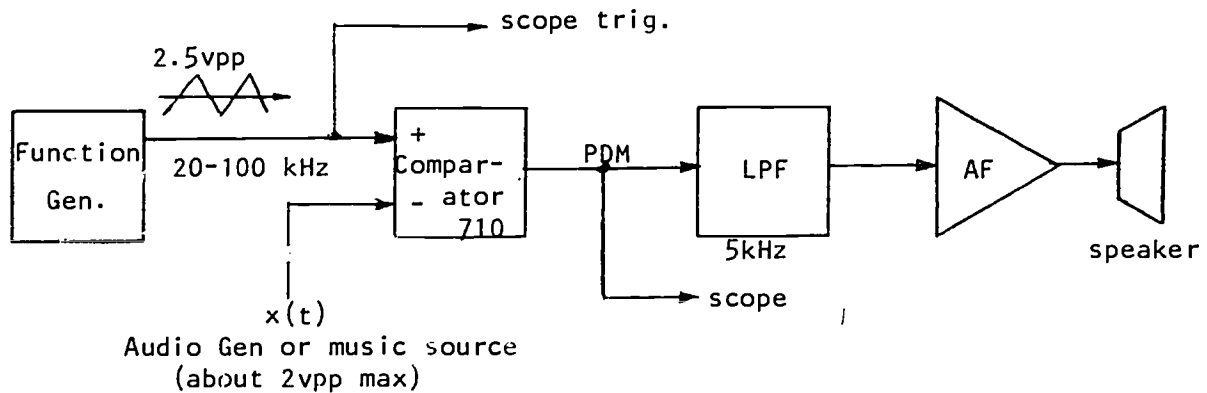
Fig. 7. PDM and reconstruction.

Analog signals are often encoded into a digital code before transmission (or further processing). This is called pulse-code-modulation (PCM). The elements of PCM generation are shown in Fig. 8.
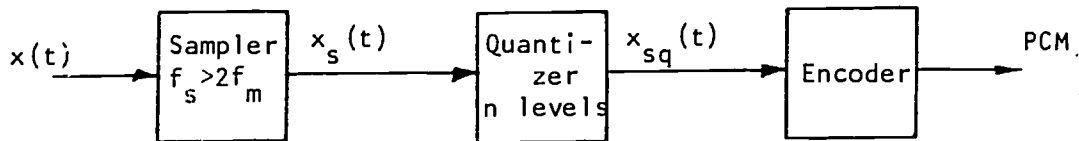
Fig. 8. PCM generation.

The analog signal $x(t)$ is first sampled to give $x_s(t)$. The sample values are then rounded off to the nearest predetermined discrete value. This is called quantization. The sampled and quantized signal is discrete both in time and in amplitude. For a finite number of levels n, each level can be represented by

a digital code of finite length. The output of the encoder forms the baseband
PCM signal. Fig. 9 illustrates these operations.
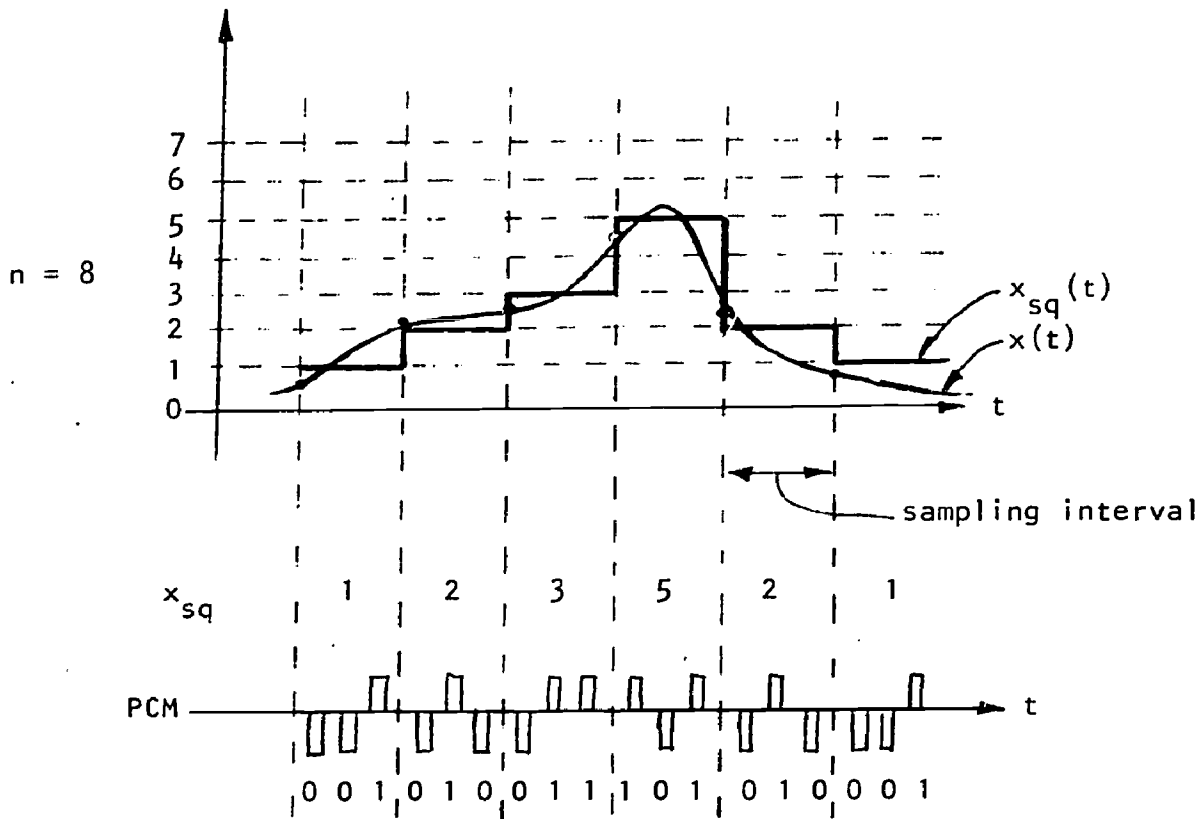
Fig. 9. Sampling, quantizing, and binary encoding.

Perfect message reconstruction is not possible in PCM systems since we are
approximating $x(t)$ by its quantized version. This quantization gives rise to
so-called quantization noise, whose value depends upon how many levels are used.
The more levels, the less noise.

The quantization of speech with variable number of levels from 2 to 256 will
be demonstrated in class with the system shown in Fig. 10. Also if time allows,
a form of PCM called "Delta modulation" will be explained and demonstrated.
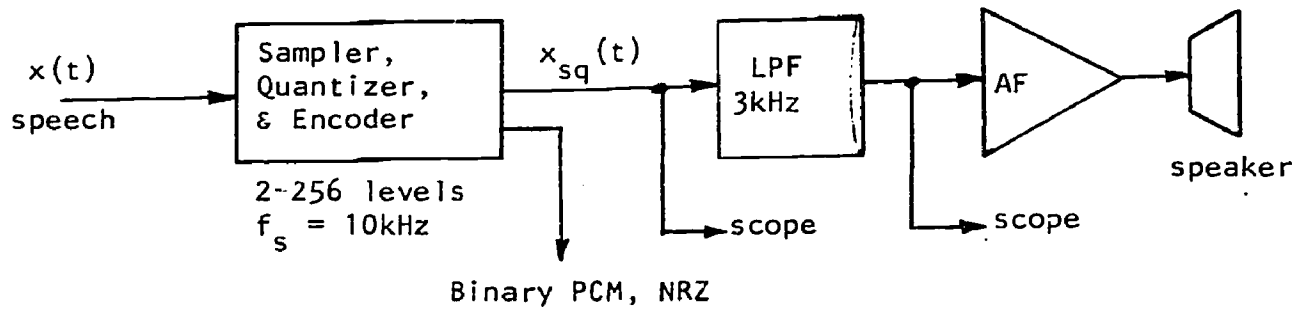
Fig. 10. Demonstration of quantization noise.

EXPERIMENT NO. 5

ANALOG MODULATION I: AMPLITUDE MODULATION,

FREQUENCY CONVERSION, AND FREQUENCY DIVISION MULTIPLEXING

Reference: A. Bruce Carlson, Chapter 5.

Introduction to Amplitude Modulation

Most baseband signals cannot be sent directly over the channel. Instead, a carrier wave is modified in accordance to the baseband signal, or modulation signal, or message, to produce a signal whose properties are better suited for the transmission medium in question. There are several types of modulation, and in this and the following experiment we will consider continuous wave (CW) modulation using analog baseband signals as the message to be transmitted. In CW modulation the carrier wave is a sinusoid. Since the ultimate goal is to send the message from one place to another, modulation must be reversable, i.e. demodulation to recover the original message must be possible.

Consider the carrier wave
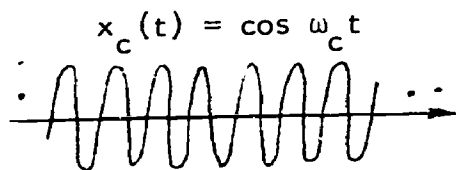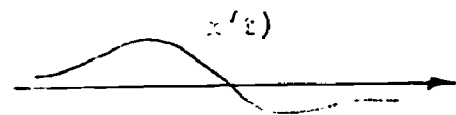
$$x_c(t) = A(t) \cos(\omega_c t + \phi(t)). \tag{1}$$

If we let the amplitude, $A(t)$, vary linearly with a modulating signal $x(t)$, the result is amplitude modulation (AM). By varying $\phi(t)$ linearly with $x(t)$ one gets phase modulation (PM), and if the time derivative of $\phi(t)$ is varied with $x(t)$ the result is frequency modulation (FM). In this experiment we will consider AM systems.

Amplitude modulation is basically a multiplication process of a high frequency carrier $x_c(t) = \cos \omega_c t$ by a message or baseband signal $x(t)$. Thus the mathematically simplest form is

$$x(t)\, x_c(t) = x(t) \cos \omega_c t \tag{2}$$

With $x(t)$ bandlimited (the usual case) it is illustrative to consider the time and frequency domain forms of (2); this is done in Fig. 1.

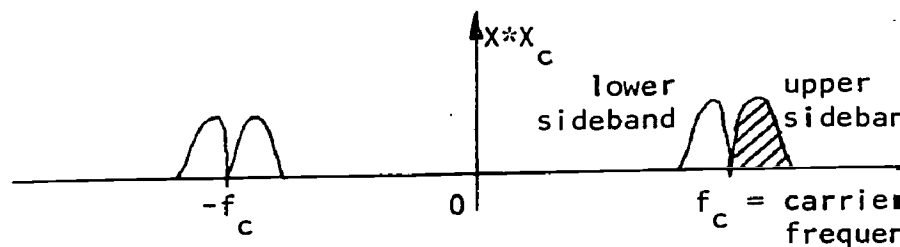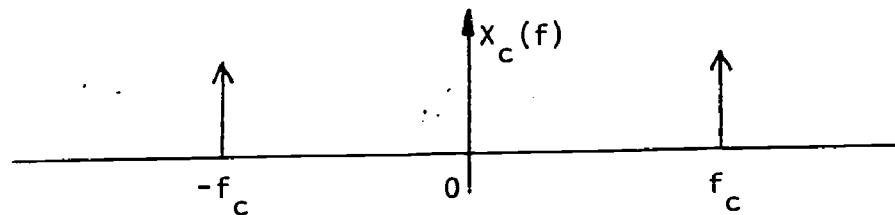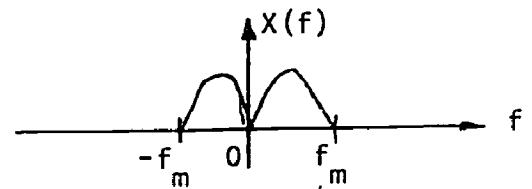Time Domain: Frequency Domain:

AM signal

Fig. 1. AM considered as a product.

Notice that with x(t) bandlimited and not containing DC, this form of AM is simply a frequency translation of the spectrum of x(t) up to the carrier frequency f_c . Furthermore, the spectrum of the AM signal does not contain the carrier itself (f_c) , consisting only of an upper sideband (shaded portion) and a lower sideband. For these reasons this form of AM is called Amplitude Modulation - Double Sideband - Suppressed Carrier (AM-DSB/SC). Historically, it was not the first to be used and it is not the form used in broadcasting as will be shown later.

To further illustrate this form, let us use a single tone for x(t) , thus with

$$x(t) = A_m \cos \omega_m t , \qquad (3)$$

the AM-DSB/SC signal becomes

$$x(t) \, x_c(t) = A_m \cos \omega_m t \, \cos \omega_c t = \frac{A_m}{2} \cos(\omega_c - \omega_m)t + \frac{A_m}{2} \cos(\omega_c + \omega_m)t . \qquad (4)$$

Equation (4) contains only two sidebands at frequencies $f_c \pm f_m$ and no carrier power. Its time domain form and amplitude spectrum is shown in Fig. 2.
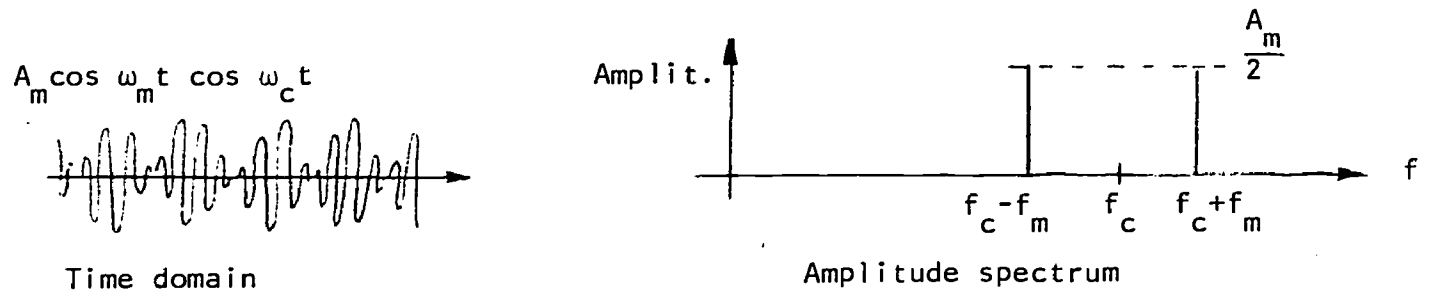
$A_m \cos \omega_m t \cos \omega_c t$

Time domain

Amplit.

$\frac{A_m}{2}$

$f_c - f_m$ f_c $f_c + f_m$ f

Amplitude spectrum

Fig. 2.   AM-DSB/SC with tone modulation.

To produce AM-DSB/SC one simply uses a multiplier, as shown in Fig. 3.

$x(t)$ $x(t) \cos \omega_c t$
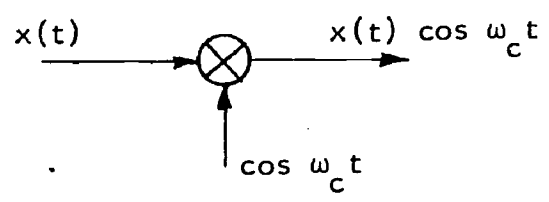
$\cos \omega_c t$

Fig. 3.   Multiplier (Mixer) for AM generation.

Since this AM signal is simply a frequency translation, then to demodulate, or recover the message, one must perform a 2nd frequency translation, which can be accomplished by a 2nd multiplier. Thus consider the product

$$\left[x(t) \cos \omega_c t \right] \cos \omega_c t = x(t) \cos^2 \omega_c t = \frac{x(t)}{2} + \frac{x(t)}{2} \cos 2\omega_c t \qquad (5)$$

As seen the first term on the left-hand side of (5) is the desired recovered message. The 2nd term is another AM-DSB/SC signal at twice the carrier frequency and is simply removed by a LPF. Fig. 4 illustrates this.
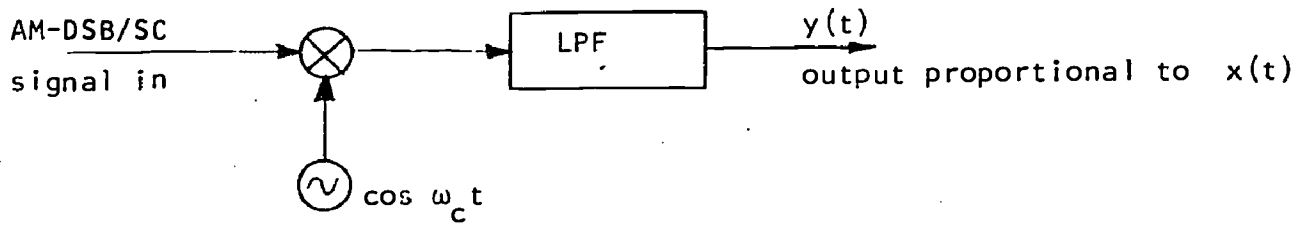
Fig. 4. Demodulation of AM-DSB/SC.

The above demodulator is called a synchronous demodulator, and requires the carrier frequency (f_c) at the receiver. The reason for this requirement is that the information (message) resides both in the amplitude and in the phase of the AM signal. This may be seen by using a switching waveform for $x(t)$; see Fig. 5.
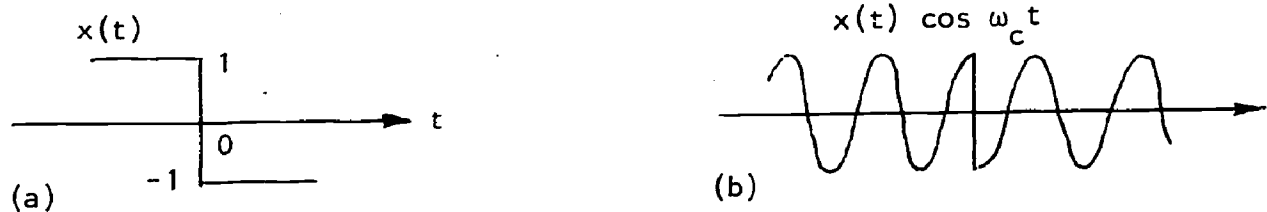

Fig. 5. AM-DSB/SC with a switching function.

At the time when $x(t)$ makes a transition from +1 to -1 the AM signal changes phase by π (180°). To extract this phase information one needs the carrier for comparison. Since the AM signal does not contain the carrier it must be generated locally (local oscillator or LO) with high precision, since small errors in frequency and phase gives rise to large errors in the demodulated signal. This is expensive if there are a large number of receivers operating, such as in broadcasting.

To avoid the LO, one must eliminate the phase shift; thus one must eliminate sign changes in $x(t)$. This is accomplished simply by adding a DC voltage of sufficient magnitude to $x(t)$ before multiplication, thus

$$\text{AM signal} = (x(t) + A) \cos\omega_c t , \quad \text{with} \quad A > |x(t)| \quad \text{all} \quad t , \tag{6}$$

yields so-called AM-DSB with a large carrier present.

Fig. 6 illustrates AM-DSB using tone modulation.

$$x(t) + A$$

(a)

(b)

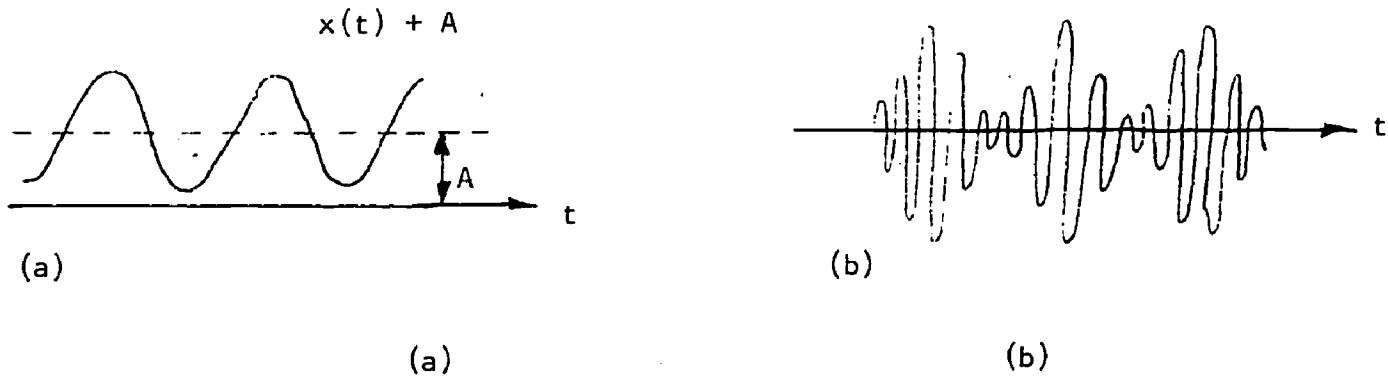(a)   (b)

Fig. 6.   AM-DSB.

From Fig. 6 it can be seen that all the information is in the envelope and thus one only needs a simple envelope detector for demodulation.   (See reference.)

The envelope detector may be considered a synchronous detector with the LO supplied by the AM signal. This is the system used for broadcasting.

Going back to the DSB/SC system, one notes that only one sideband (either the upper or the lower) is sufficient for transmitting the information about $x(t)$. Thus, the last AM system uses only one sideband instead of two and is called single-sideband/suppressed carrier (AM-SSB/SC). There are several ways to generate SSB/SC; the one we will use simply uses a filter to remove one of the sidebands. Demodulation is again accomplished by a synchronous detector requiring the carrier frequency. For example, using tone modulation the SSB/SC signal is simply one term of (4):

$$SSB/SC_{upper} = \frac{A_m}{2} \cos(\omega_c + \omega_m)t \ . \tag{7}$$

Problem

Show that multiplication by $\cos \omega_c t$ followed by LPF will demodulate (7).

Errors

Since the LO will never be exactly correct in frequency and phase, there will be errors in the demodulated signal. We will consider the errors here for SC systems using tone modulation and the demodulator of Fig. 4.

I.   Phase error in the LO, i.e. $LO = \cos(\omega_c t + \phi)$.

(a) DSB/SC. Demodulator output will be (show this):

$$y(t) = \frac{A_m}{2} \cos \phi \cos \omega_m t \tag{8}$$

Notice that the output level depends upon ϕ.

(Output is zero if $\phi = 90°$.) Conclusion: DSB/SC is phase sensitive.

(b) SSB/SC, then (show this)

$$y(t) = \frac{A_m}{4} \cos(\omega_m t + \phi) \tag{9}$$

The phase error in $y(t)$ is not audible to the human ear, thus one concludes that for speech or music the SSB/SC is not phase sensitive.

II. Frequency error in the LO, i.e.

$$LO = \cos(\omega_c + \Delta\omega)t = \cos(\omega_c t + \Delta\omega t) ,$$

which is equivalent to a phase error $\phi = \Delta\omega t$, thus

(a) DSB/SC $\qquad y(t) = \frac{A_m}{2} \cos \Delta\omega t \cos \omega_m t \tag{10}$

If $\Delta f = \Delta\omega/2\pi$ is very small, say 1 Hz or less this will sound like a slow variation of the tone (f_m) level. If Δf is larger, (10) may be expanded to

$$y(t) = \frac{A_m}{4} \left[\cos(\omega_m - \Delta\omega)t + \cos(\omega_m + \Delta\omega)t\right]$$

which gives two tones, none of which is the correct (f_m).

(b) SSB/SC $\qquad y(t) = \frac{A_m}{4} \cos(\omega_m + \Delta\omega)t , \tag{11}$

thus we get a change of tne pitch by Δf.

Conclusions: Both SSB/SC and DSB/SC are frequency sensitive. For speech
the effect on SSB/SC is less serious than that on DSB/SC.
Why?

Procedures

I. Modulation.

Generate the various forms of AM using the block diagram of Fig. 7. Use tone modulation at about 1-5 kHz and observe the signals both in time and frequency.
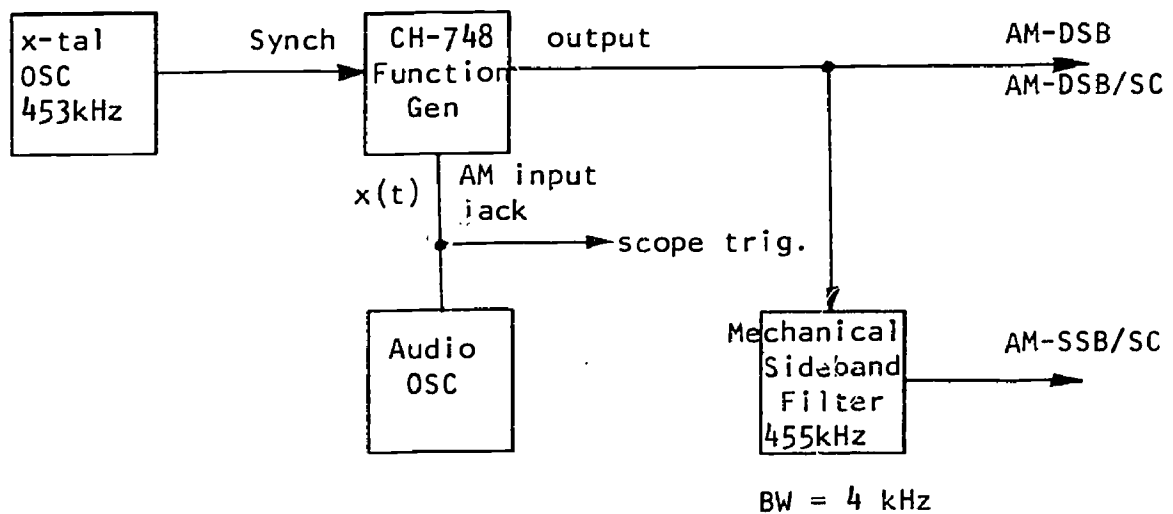
Fig. 7. AM modulation.

The x-tal oscillator sets the carrier frequency of the CH-748 to the edge of the sideband filter. Before attempting to modulate make sure that the CH-748 is locked to the x-tal oscillator. Locking may be detected by triggering a scope from the x-tal oscillator while observing the output of the CH-748. A stationary display indicates lock. A single x-tal oscillator is used for all squads; its output is available from the lab distribution system. Be sure not to overdrive the filter (it has gain).

II. Demodulation.

(a) Using an HP10534A mixer as a multiplier set up a synchronous demodulator as in Fig. 8 for the SC types of AM.
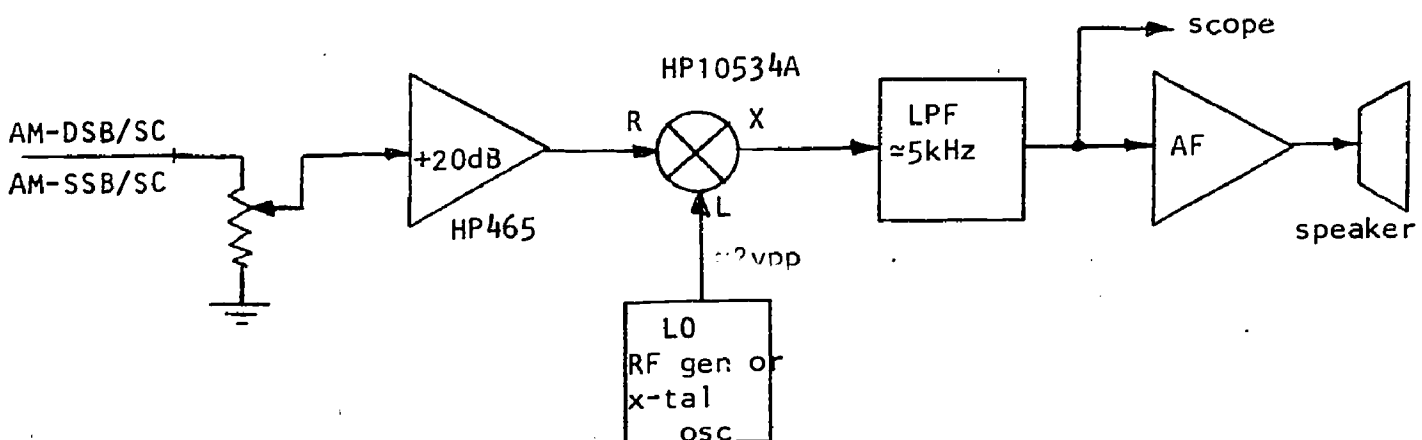

Fig. 8. Demodulation of AM-DSB/SC and AM-SSB/SC.

For the LO, try both a RF generator (LG-21) and a variable frequency x-tal oscillator. Use music and/or speech as a modulating signal and listen to the demodulated signal while trimming the LO frequency. Do this for both DSB/SC and SSB/SC. Does the low frequency (base) response of the two systems differ? If so, why? Any conclusions?

(b) Use an envelope detector to attempt to demodulate AM-DSB/SC modulated with music. Slowly add DC (internal to 748) to $x(t)$ until you have AM-DSB. Conclusions?
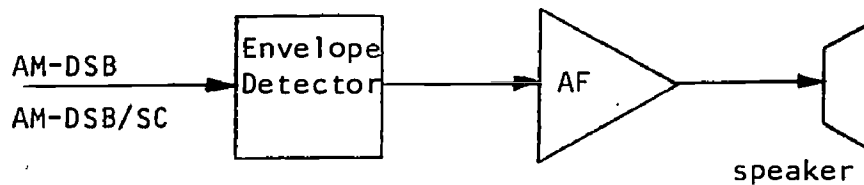
Fig. 9. Envelope detection.

Frequency conversion.

Amplitude modulation and demodulation are essentially frequency translations. Such translations are often called frequency conversion when they are not used specifically for modulation or demodulation. A multiplier (mixer) may be used for general frequency conversion. Consider Fig. 10.
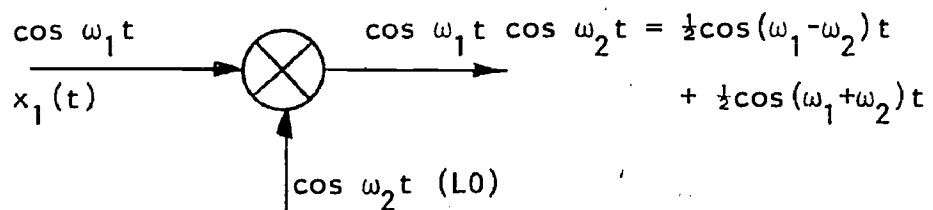
Fig. 10. Frequency conversion.

Note that the input frequency f, has been changed into two frequencies $|f_1 \pm f_2|$. If we extract $|f_1 - f_2|$ by a suitable filter we have a downconverter; by extracting $f_1 + f_2$ we get an upconversion of the input. The spectrum of a general input $x_1(t)$ may be translated up or down as needed by the appropriate choice of the LO frequency and the output filter (called an intermediate amplifier (IF)).

Demonstrate frequency conversion using the block diagram of Fig. 11. What frequency components do you expect to get out?
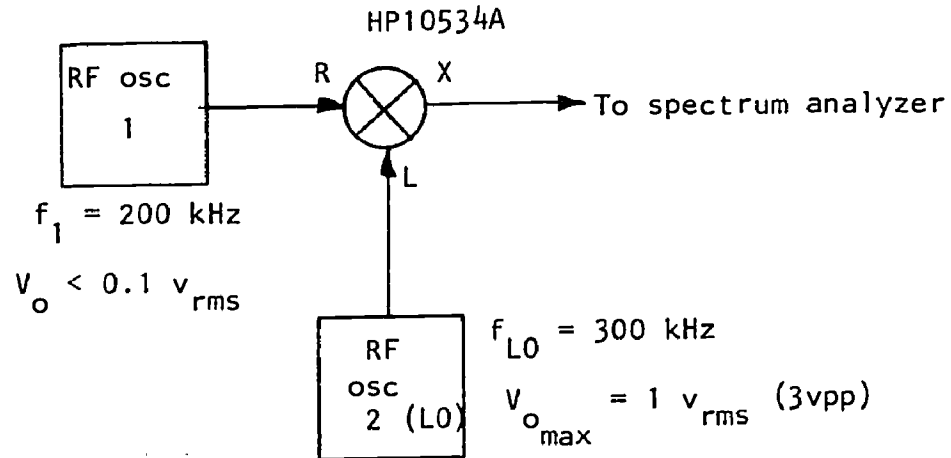
HP10534A

```
+----------+        R       X
| RF osc   |------>( X )------> To spectrum analyzer
|   1      |         |
+----------+         | L
                     |
f_1 = 200 kHz        |
                     |
V_o < 0.1 v_rms      |
              +----------+   f_LO = 300 kHz
              |   RF     |
              |  osc     |   V_o_max = 1 v_rms  (3vpp)
              |  2 (LO)  |
              +----------+
```

Fig. 11. Frequency conversion.

Frequency Division Multiplexing (FDM).

By allocating channels in the spectrum several channels may share the same transmission medium. This is called FDM. In FDM the various signals exist at the same time while being separated in frequency. As you may recall in TDM the signals are separated in time while sharing the same frequency slot. FDM is implemented by frequency translating the various baseband signals to their assigned locations. Radio broadcasting or TV are examples of FDM.

Thus one sees that amplitude modulation, frequency conversion, and frequency division multiplexing are all related operations.

Vestigial-Sideband Modulation (VSB).

Bandwidth conservation argues for the use of SSB, but practical SSB systems have poor low-frequency response. DSB has good low-frequency response but the transmission bandwidth is twice that of SSB. A compromise called VSB has good low-frequency response and has a bandwidth just a little bit larger than SSB. For details of VSB see reference, page 198.

ANALOG MODULATION II: FREQUENCY AND PHASE MODULATION

Reference: A. Bruce Carlson, Chapter 6.

Introduction

As seen in Experiment 5 amplitude modulation is basically a frequency translation whose bandwidth never exceeds twice the message bandwidth. Another property is that the demodulated signal-to-noise ratio is no better than baseband transmission (direct transmission without modulation) and can only be improved by increasing the transmitter power.

In frequency and phase modulation (also called angle modulation) the modulated spectrum is not related in a direct fashion to the message spectrum and is usually much greater than twice the message spectrum. A consequence of this is that angle modulation can provide increased signal-to-noise ratios without increased transmitter power.

In Experiment 5 we pointed out that there are other ways besides AM which we could modulate a sinewave in accordance with some message signal. Before going on we need first specify what we mean by frequency (or phase) modulation and the idea of instantaneous frequency.

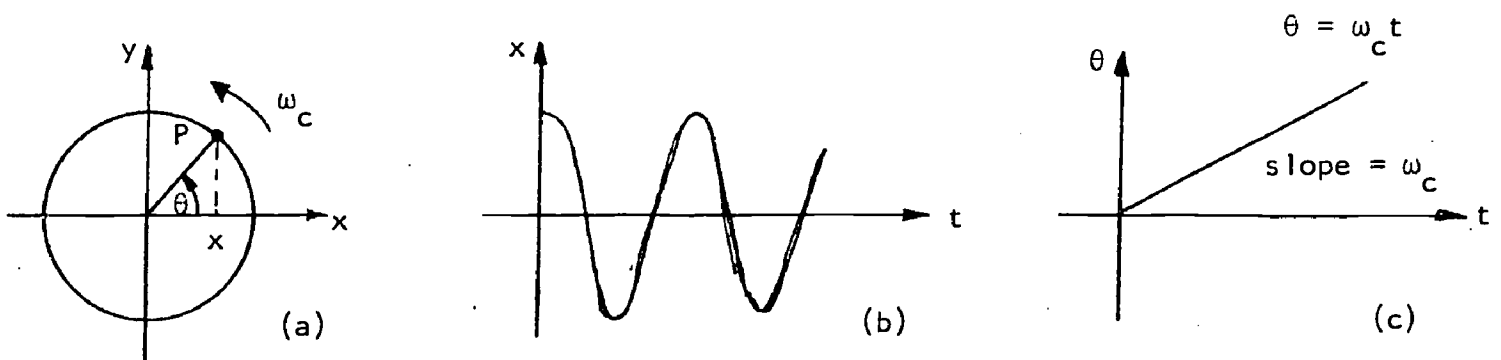Consider a rotating wheel of unit radius, Fig. 1.

Fig. 1. Rotating wheel and the motion of the projection of P on the x-axis.

39

If the angular velocity ω is a constant then the angle θ increases linearly with time, i.e. $\theta = \omega_c t$, (c); and the projection of point P on the x-axis, $x = \cos\theta$, describes harmonic motion as in (b). In this case we may talk about the frequency, $f_c = \omega_c/2\pi$, of the wheel.

The angle θ does not need to be a linear function of t however; in fact, the wheel may speed up or slow down, or alternate between acceleration and deaccel- eration, Fig. 2.

Fig. 2. Non-linear variation of θ with t.

In Fig. 1(c) we note that the angular velocity is the slope of the straight line variation of θ, thus for the non-linear variation of Fig. 2 we may define an instan- taneous angular velocity $\omega_i(t)$ as the slope at any point of the curve

$$\omega_i(t) = \frac{d\theta(t)}{dt} \quad . \tag{1}$$

Note that $\omega_i(t)$ is a function of time.

The non-linear variation in Fig. 2 may be thought of as consisting of a linear part added to a non-linear part

$$\theta(t) = \omega_c t + \phi(t) , \tag{2}$$

thus using (1) and (2), the instantaneous angular velocity becomes

$$\omega_i(t) = \omega_c + \frac{d\phi(t)}{dt} , \tag{3}$$

and the general motion of the projection of P on the x-axis may be written

$$x = \cos\theta(t) = \cos(\omega_c t + \phi(t)) . \tag{4}$$

40

With $\phi(t) = 0$ or a constant we get the simple harmonic motion (sinewave) of Fig. 1. The whole point of this boils down to the fact that we can only take the cosine (or sine) of an angle and that we must distinguish between instantaneous frequency and Fourier (or spectral) frequency which implies periodicity.

Now let us apply these ideas to angle modulation, and consider (4) a general signal

$$x_c(t) = \cos \theta(t) = \cos(\omega_c t + \phi(t)) \tag{5}$$

with $f_c = \omega_c/2\pi$ the carrier frequency and $\phi(t)$ the phase angle.

Phase Modulation (PM)

If we let the phase angle, $\phi(t)$, be a linear function of a modulating signal $x(t)$, the result is PM. Thus

$$\phi(t) = k_\phi x(t) , \tag{6}$$

where k_ϕ is a constant called the modulation sensitivity (in say radians/volts).

The PM signal is simply the cosine of the angle $\theta(t)$, thus

$$\text{PM signal} = \cos \theta(t) = \cos(\omega_c t + k_\phi x(t)) , \tag{7}$$

with a carrier frequency $f_c = \omega_c/2\pi$ and instantaneous frequency

$$f_i(t) = \frac{\omega_i}{2\pi} = \frac{1}{2\pi} \frac{d\theta(t)}{dt} = f_c + \frac{k_\phi}{2\pi} \frac{dx(t)}{dt} . \tag{8}$$

Frequency Modulation (FM)

In FM we let the instantaneous frequency be a linear function of the modulation signal, thus

$$f_i(t) = f_c + k_f x(t) , \tag{9}$$

when k_f is the modulation sensitivity for FM (say in kHz/volts). To find the FM signal we need the angle $\theta(t)$.

From (1)

$$\theta(t) = \int \omega_i(t)dt = 2\pi \int f_i(t)dt \ , \tag{10}$$

and using (9) in (10), we get

$$\theta(t) = \omega_c t + 2\pi k_f \int x(t)dt \ , \tag{11}$$

finally

$$FM \ signal = \cos \theta(t) = \cos(\omega_c t + 2\pi k_f \int x(t)dt) \ . \tag{12}$$

In order to get a better feeling for PM and FM and their differences, let us use a specific form of $x(t)$ and find the time domain forms of the resulting PM and FM signals. Fig. 3 illustrates the resulting forms. If $x(t)$ is a ramp, $x(t) = a$ then the FM instantaneous frequency increases linearly with time, i.e. $f_i = f_c + k_f$ while for PM there occurs a frequency shift at $t = 0$. This may be seen using (8).
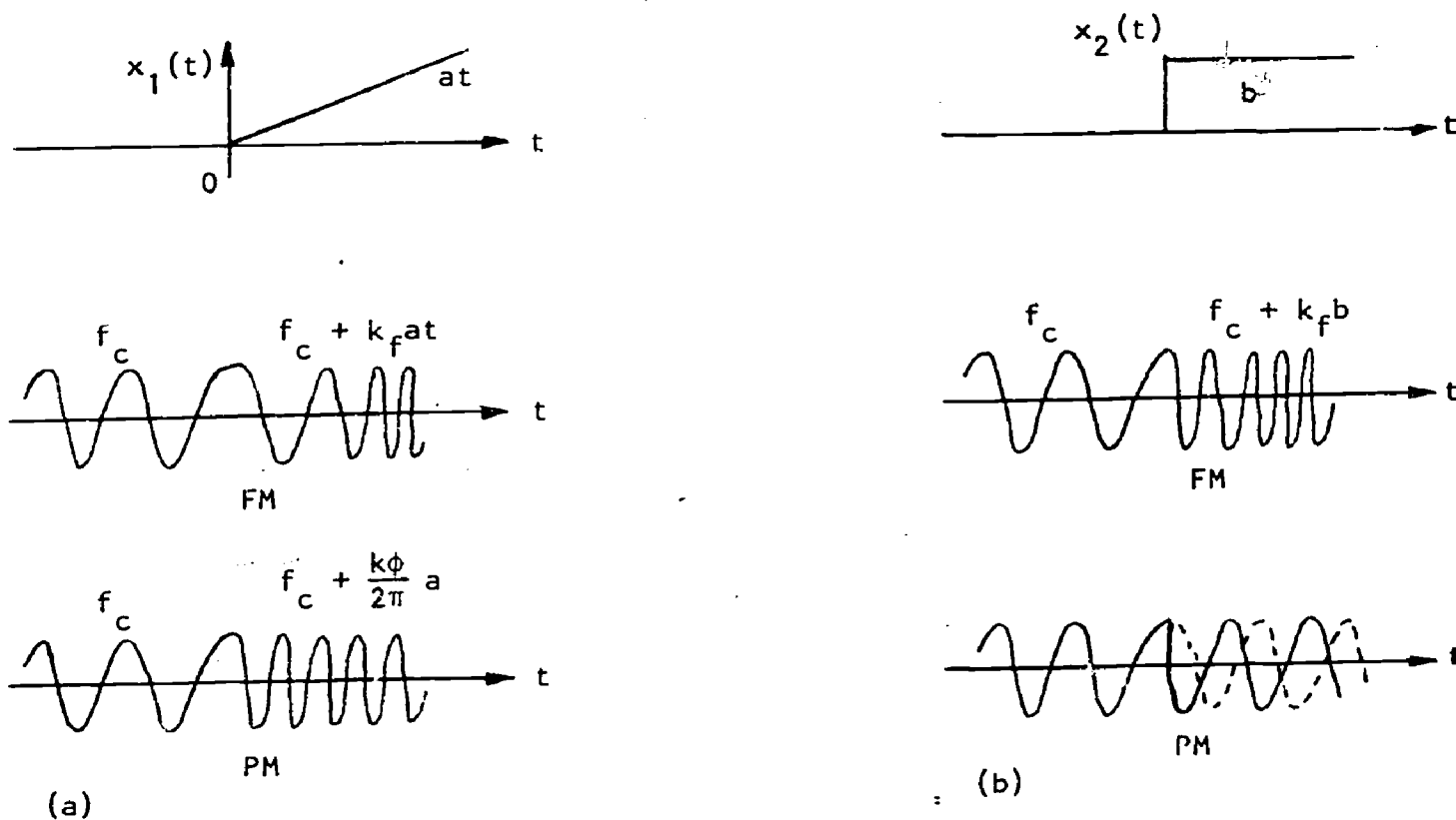
Fig. 3. FM and PM waveforms for (a), a ramp; and (b) a step modulating signal.

In (b) $x(t)$ is a step. In this case, there is a sudden frequency shift at $t = 0$ in the FM signal, while for PM there occurs a sudden shift in the phase angle. Note that PM modulated with a ramp looks the same as FM modulated with a step. The reason is of course that $x_1(t)$ is the integral of $x_2(t)$. Both FM and PM have time varying phase and frequency, and are quite similar. In fact, for tone modulation it is not possible to visually distinguish FM and PM waves. Equations (6) through (12) indicate that with the help of differentiating and integrating networks, a frequency modulator can produce phase modulation and vice versa.

In the following we will concentrate on FM.

Consider a tone modulating signal

$$x(t) = A_m \cos \omega_m t , \tag{13}$$

then the instantaneous frequency is from (9)

$$f_i = f_c + k_f A_m \cos \omega_m t = f_c + \Delta f \cos \omega_m t \tag{14}$$

where $\Delta f = k_f A_m$ is the maximum frequency deviation.
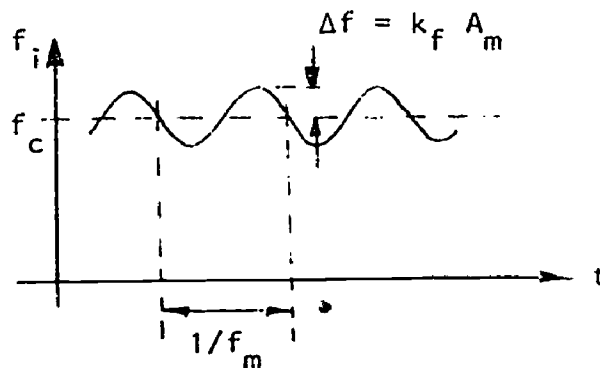
Equation (14) is sketched in Fig. 4.

Fig. 4.   Instantaneous frequency vs. time for
tone modulation.

Note that the rate at which the instantaneous frequency varies is the same as the tone frequency (f_m) and the maximum excursions (Δf) is proportional to the tone level (A_m).

$\theta(t)$ is obtained from (11),

$$\theta(t) = \omega_c t + \frac{2\pi k_f A_m}{\omega_m} \sin \omega_m t = \omega_c t + \frac{\Delta f}{f_m} \sin \omega_m t ,$$

and letting $\beta = \dfrac{\Delta f}{f_m}$, the modulation index for tone modulated FM,

we get

$$\theta(t) = \omega_c t + \beta \sin \omega_m t \ .$$ (15)

Thus the FM signal is

$$x_{FM}(t) = \cos(\omega_c t + \beta \sin \omega_m t) \ .$$ (16)

(16) is very nearly periodic if $\omega_c \gg \omega_m$, and may therefore be expanded into a Fourier series. Expansion yields (see reference)

$$
\begin{aligned}
x_{FM}(t) = \ &J_0(\beta)\cos \omega_c t - J_1(\beta)\Big[\cos(\omega_c - \omega_m)t - \cos(\omega_c + \omega_m)t\Big] \\
&+ J_2(\beta)\Big[\cos(\omega_c - 2\omega_m)t + \cos(\omega_c + 2\omega_m)t\Big] \\
&- J_3(\beta)\Big[\cos(\omega_c - 3\omega_m)t - \cos(\omega_c + 3\omega_m)t\Big] \\
&+ \ . \ . \ . \ .
\end{aligned}
$$ (17)

where $J_n(\beta)$ = Bessel function of the first kind, index n, and argument β . (For details of the Bessel functions, see reference.)

(17) is a time function consisting of a carrier and an infinite number of side-bands spaced at $\pm f_m$, $\pm 2f_m$, etc., away from the carrier. A typical spectrum is shown in Fig. 5.
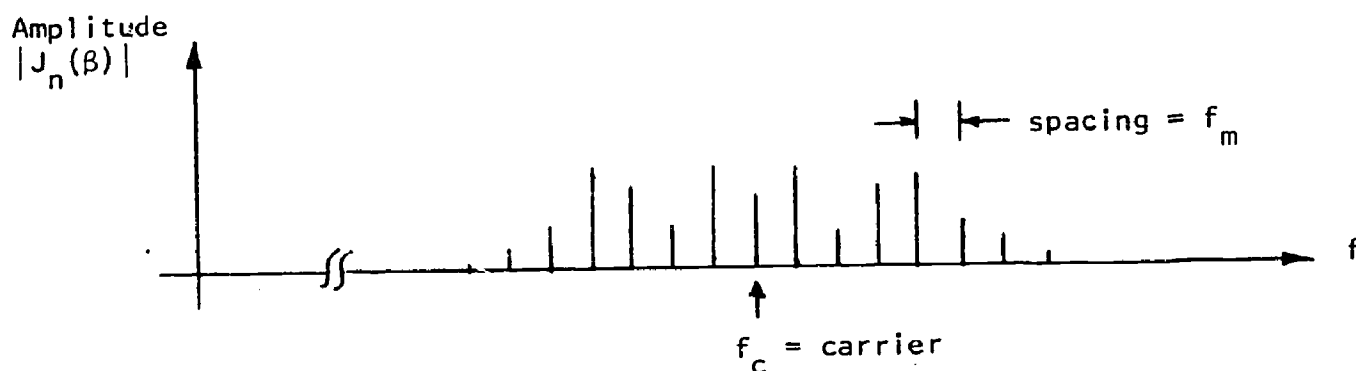

Fig. 5. FM spectrum for $\beta = 5$.

This is a wideband signal as contrasted to AM. Furthermore the magnitudes of the carrier and the sidebands depend upon the modulation index β . This is again

different from AM-DSB where the carrier amplitude is fixed.

The BW of the FM signal depends upon the number of significant sidebands. A sideband is significant if its amplitude is greater than or equal to 0.01 (1% of unmodulated carrier). To determine BW in this manner requires a table of Bessel functions and knowledge of β . A more convenient rule will be derived in class and is given here

$$BW_{FM} = 2(\Delta f + 2f_m) . \tag{18}$$

(18) is sufficiently accurate for $\beta > 2$.

For small values of β , say $\beta \ll 1$, only the two sidebands next to the carrier are significant, and the BW is then the same as AM (i.e. $2f_m$). This is called Narrow-Band-FM (NBFM). The reader is referred to the reference for further details.

Procedures

I. Generation of FM

The generation of FM may be implemented in several ways, one of which is the use of a Voltage-Controlled-Oscillator (VCO). The output of such an oscillator is a sinewave (or a squarewave) whose frequency is proportional to a control voltage. Fig. 6 shows a VCO (CH-748 AM/FM function generator) and its characteristic.
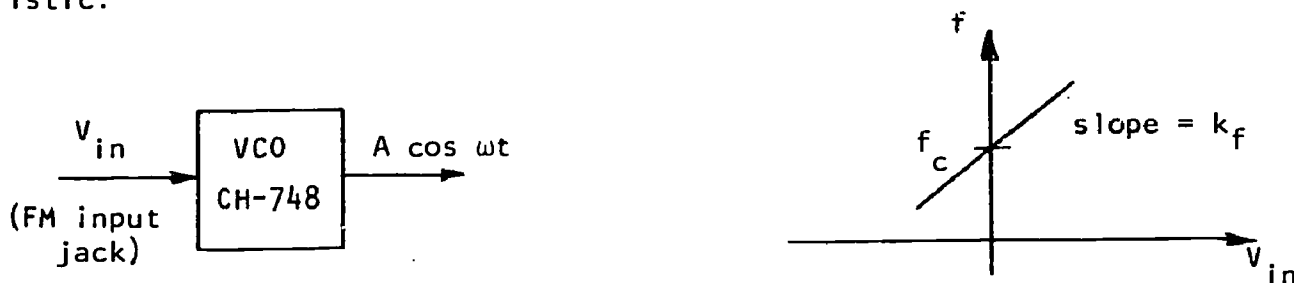

Fig. 6. VCO and its characteristic.

a) Set the free running ($V_{in} = 0$) frequency of the CH-748 at about 450 kHz, and then measure the frequency of oscillation for two different input voltages (say $V_{in} = 0$, and 0.2 volts) using a counter. Compute the modulation sensitivity, k_f , from these measurements.

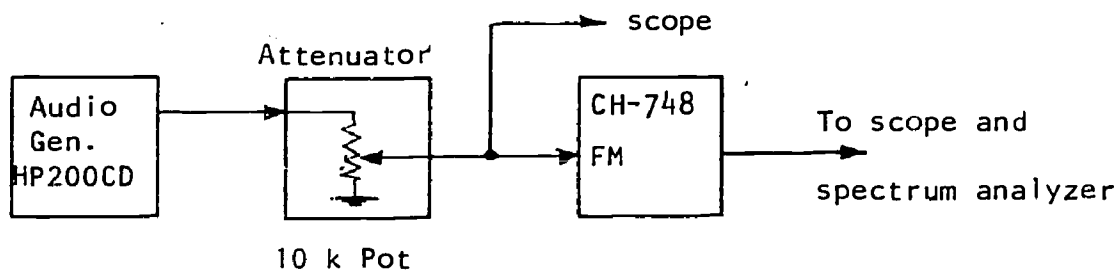b) Using the block diagram of Fig. 7 generate a tone modulated FM signal.

Fig. 7. FM tone modulation.

First observe the FM signal in the time domain using a low modulation frequency (5 Hz or less) starting with zero modulation and slowly increase the audio level. For a given level make changes in the audio frequency in order to understand its effect.

Second, change the audio frequency to about 5 kHz and observe the FM spectrum on the spectrum analyzer. Again start with zero modulation and slowly increase the audio level. Adjust the spectrum analyzer so that the carrier is displayed in the middle of the sweep, and use a dispersion of 10 kHz/cm so as to be able to see the individual sidebands. Note that without modulation (audio level = zero) only a carrier is present. As the modulation is increased the sidebands appear and the carrier level decreases. At some value of audio level the carrier vanishes. Measure the audio amplitude (and note the audio frequency) at which the carrier disappears. At this point $\beta = 2.405$ (the first zero of $J_0(\beta)$). From these data calculate k_f.

This is called the carrier null method for measuring the modulation sensitivity. Estimate the FM BW at this point and compare to (18).

II. Demodulation of FM

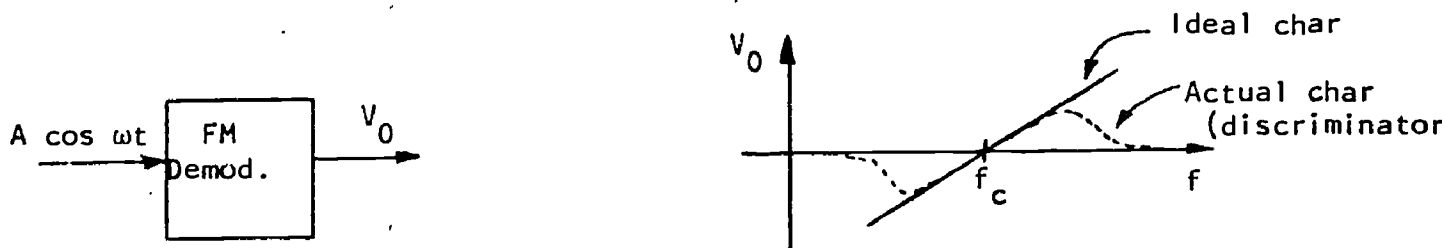To demodulate FM one needs a device such as that illustrated in Fig. 8.

Fig. 8. FM demodulator.

There are several ways to implement the characteristic shown in Fig. 8. The actual characteristic in the figure is typical of so-called discriminators, and is usually called an S-curve. A better fit to the ideal can be obtained by using a phase-locked-loop (PLL).*

Use the CH-748 VCO as a sweep generator for the dynamic display of the discriminator S-curve, Fig. 9.
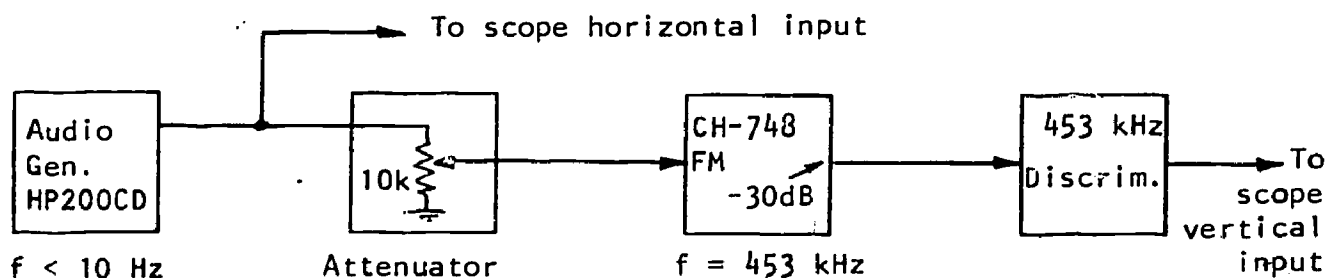
To scope horizontal input

| Audio Gen. HP200CD | | 10k (Attenuator) | CH-748 FM -30dB | 453 kHz Discrim. | To scope vertical input |

f < 10 Hz   Attenuator   f = 453 kHz

Fig. 9.   Scope display of discriminator S-curve.

II.   Complete FM system

Set up a complete FM system as in Fig. 10.

From music source — 10k Pot — CH-748 VCO -30dB — Discrim. 453 kHz — AF — speaker
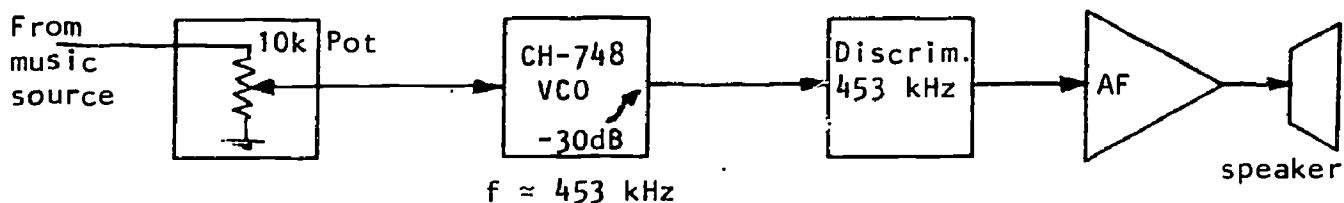
f ≈ 453 kHz

Fig. 10.   Complete FM system.

Using a music or speech signal, observe (listen to) the effect of VCO tuning and frequency deviation, Δf , (adjusted by the level pot) on the output from the speaker. Notice that as Δf is increased the output level increases without changing the "power transmitted." This is the reason for the better

*A. B. Grebene: "The monolitic phase-locked-loop, a versatile building block," IEEE Spectrum, Vol. 8, No. 3, pp. 38-49 (March 1971).

signal-to-noise ratio of the FM system over that of AM (as long as the FM signal is above a certain threshold). To increase the signal-to-noise ratio in an AM system one must increase transmitted power. We will come back to this later when we test an FM receiver. What happens when $2\Delta f$ becomes larger than the linear portion of the S-curve?

IV. Generation of PM

The characteristics of the CH-748 is such that it will produce phase modulation when modulated at the FM jack and locked to an external reference at the carrier frequency.

a) Generate a tone modulated PM signal using the set-up of Fig. 11.
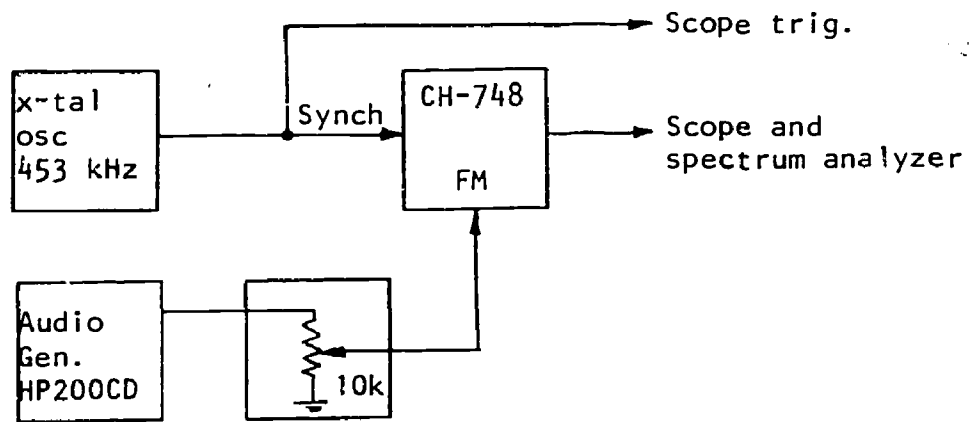
Fig. 11. PM generation.

For tone modulation, $x(t) = A_m \cos \omega_m t$, the PM signal is according to (7)

$$x_{PM}(t) = \cos(\omega_c t + k_\phi A_m \cos \omega_m t) . \tag{19}$$

Letting $k_\phi A_m = \Delta\phi$ ($= \beta$ of FM) the maximum phase deviation, (19) may be written

$$x_{PM}(t) = \cos(\omega_c t + \Delta\phi \cos \omega_m t) , \tag{20}$$

which is of the same form as tone modulated FM.

Before applying any modulating signal make sure that the CH-748 generator is locked to the x-tal oscillator. When locking has been achieved, modulate using an audio tone at about 5 Hz or less and observe the time domain form

48

of the PM signal. Vary both the audio level and frequency to understand their effects. (The audio level must be restricted to avoid losing lock; keep $\Delta\phi = \beta <$ about $\pi/4$, i.e. less than about 45°.)

Next go to about 5 kHz modulating frequency and observe the spectrum of the PM signal for various audio levels and frequencies, again making sure that lock is not lost. How does the spectrum compare to that of FM?

b) To get a better feel for FM and PM (and commercial broadcast FM) let us try to demodulate PM with an FM demodulator. Obviously, to demodulate PM with an FM demodulator is going to introduce some kind of errors. But what kind of errors? For example, how would music sound with such a system? Let us find out by considering the system in Fig. 12.
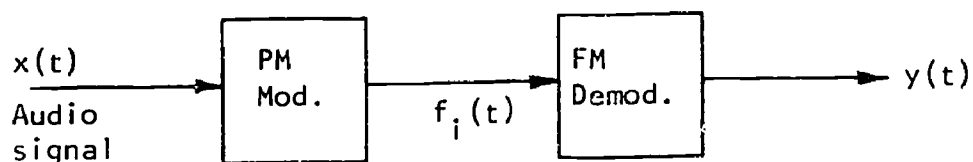

Fig. 12.

Analysis:

The FM demodulator responds to the instantaneous frequency in a linear manner, thus we need $f_i(t)$ of the PM signal.

From (7), $\theta(t) = \omega_c t + k_\phi x(t)$, thus

$$f_i(t) = \frac{1}{2\pi} \frac{d\theta(t)}{dt} = f_c + \frac{k_\phi}{2\pi} \frac{dx(t)}{dt} ,$$

and the output is proportional to $f_i(t) - f_c$, thus

$$y(t) = k_d \frac{dx(t)}{dt} , \qquad (21)$$

(k_d = constant of the demodulator).

Consider tone modulation, $x(t) = A_m \cos \omega_m t$. Then using (21), the output is

$$y(t) = k_d A_m \omega_m \sin \omega_m t . \qquad (22)$$

Note that the output level is proportional to the tone frequency. With music one would lose the base and increase the treble; it would sound very tinny. To compensate, one could integrate the output (Eqn. (21)), or use a RC-LPF on y(t) with the low-frequency cut off less than the lowest audio frequency to be sent. Such a filter has exactly the correct fall off (-6 dB/octave) to compensate for the increase in level with frequency.

Broadcast FM uses a preemphasis network ahead of the FM modulator. This network acts as a differentiator for the treble tones; thus for treble tones the transmission is PM, while for base tones it is FM. To compensate one uses a deemphasis network (RC-LPF) at the receiver. Emphasis improves the signal-to-noise ratio since speech and music contain very little power above 1 kHz. Emphasis is also used in tape and disk recordings.

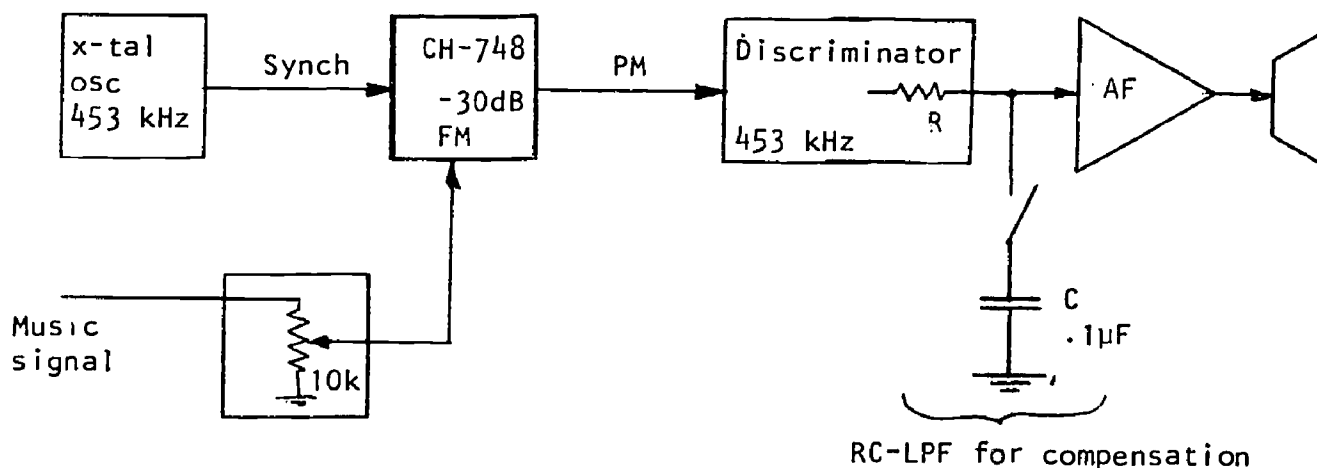Set up the PM modulator/FM demodulator of Fig. 13.

Fig. 13.   FM demodulator on a PM signal.

Listen to the reproduction without the compensating network, then add the capacitor C which with the resistor R internal to the discriminator forms the RC-LPF.

EXPERIMENT NO. 7

CHARACTERISTICS OF NOISE AND NOISE FIGURE

Reference: A. Bruce Carlson, Chapter 3.

Part 1. Characteristics of Noise.

Introduction

Unwanted electrical signals come from a variety of sources and are usually termed noise. There are man-made and naturally occurring interferences. Here we are going to be concerned with unavoidable sources of naturally occurring noise which set a fundamental limit to systems performance. One such type of noise is called thermal noise since it is due to the thermal motions of electrons in conducting media. Other noise sources are non-thermal in the sense that the noise power is unrelated to a physical temperature. An example of the latter is the so-called shot noise of a vacuum tube which is due to the random emissions of electrons from a hot cathode (shot noise also exists in semiconductors).

In dealing with random phenomena such as noise it is convenient to use the idea of spectral density which is a measure of the average power per unit frequency (or bandwidth) as a function of frequency. (This is very similar to mass density; for example, consider a thin string whose mass is a function of position. To characterize the string one could plot the mass per unit length vs. position along the string.)

Spectral density may be given in watts/Hz, $volts^2$/Hz, or $amperes^2$/Hz depending upon the application or use.

Thermal noise

When a resistor (or any conductor having resistance) R is at temperature T, there is a random voltage produced at its terminals by the random motions of the electrons in the resistor. The spectral density of this noise voltage is

$$G_v(f) = 4kTR \text{ volts}^2/\text{Hz}, \quad f \geq 0 \tag{1}$$

where k = Boltzmann's constant.

Equation (1) is valid as long as the frequency, f, is less than about 10^{12} Hz which is in the infrared portion of the electromagnetic spectrum. Thus $G_v(f)$ is essentially constant for conventional electrical devices. (1) is sketched in Fig. 1.
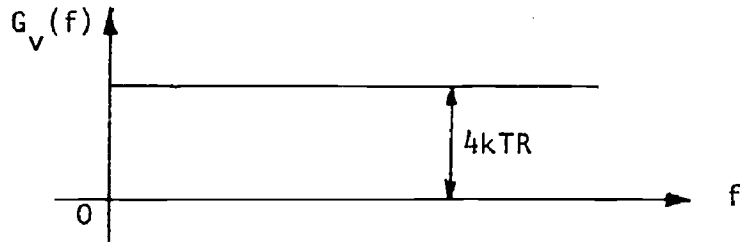

Fig. 1. Noise spectral density for a resistor.

For room temperature $T_0 = 290°K$ (17°C), $kT_0 = 4 \times 10^{-21}$ joules (watt-seconds). We may use (1) to construct a model of a noisy resistor; this is done in Fig. 2.
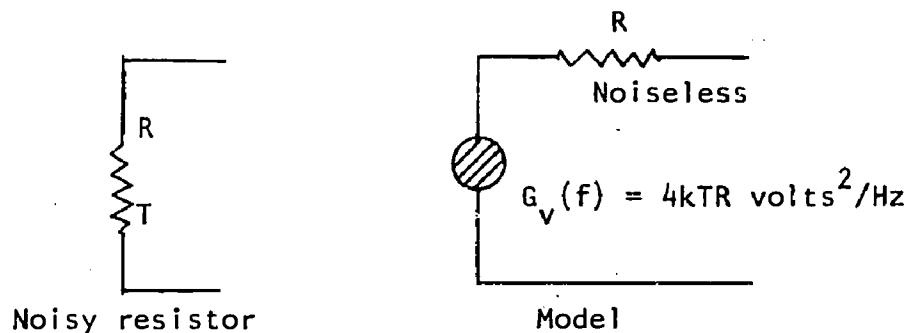
Fig. 2. Noise model of a resistor.

A condition usually desired in communication systems is that a load is matched to a source, i.e. the load resistor is equal to the source resistor. This results in maximum power transfer from the source to the load. Consider Fig. 3, where we have a matched load to a thermal noise source.
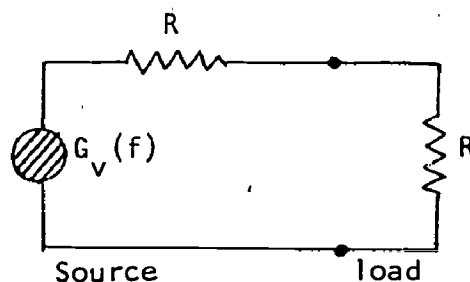

Fig. 3. Thermal resistance with matched load.

The available spectral density at the load resistance is

$$G_a(f) = \underbrace{\frac{G_v(f)}{2R}}_{\substack{\text{"Total} \\ \text{power"} \\ \text{in 2R}}} \underbrace{\frac{1}{2}}_{\substack{\frac{1}{2}\text{ power} \\ \text{in each R}}} = \frac{G_v(f)}{4R} = kT \text{ watts/Hz .} \qquad (2)$$

Thus a thermal resistor delivers kT watts/Hz to a matched load independent of the resistance value.

Besides thermal noise, many other noise sources have a flat spectral density over a wide range of frequencies. Such spectral densities are called white. Many white noise sources are non-thermal in the sense that the noise power is unrelated to a physical temperature. We can, however, always speak of a noise temperature, T_N , of a white noise source, such that its available spectral density is given by kT_N .

Procedure

(a) Observe the output of the GR1390 noise generator in both the time and the frequency domains for the three bandwidth (BW) settings of the noise generator. Estimate the upper noise cut-off frequency in each case.

(b) Use a <u>true</u> RMS voltmeter together with a Krohn-Hite variable filter to show that noise power is proportional to BW. Plot the square of the RMS voltage vs. BW of the LPF. Fig. 4.
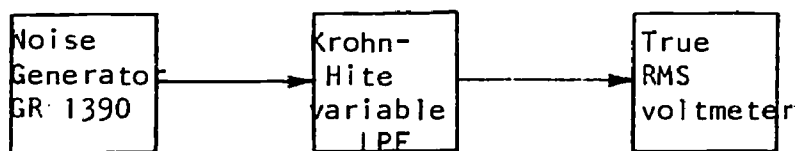

Fig. 4

(c) Using two noise sources and an op-amp (x1) summer, show that noise powers add if the two sources are incoherent, but that the noise voltages add if the two sources are coherent, Fig. 5. Repeat for sinusoidal generators. What happens as the frequencies of the two generators are brought very close together? Be sure not to overdrive the op-amp.
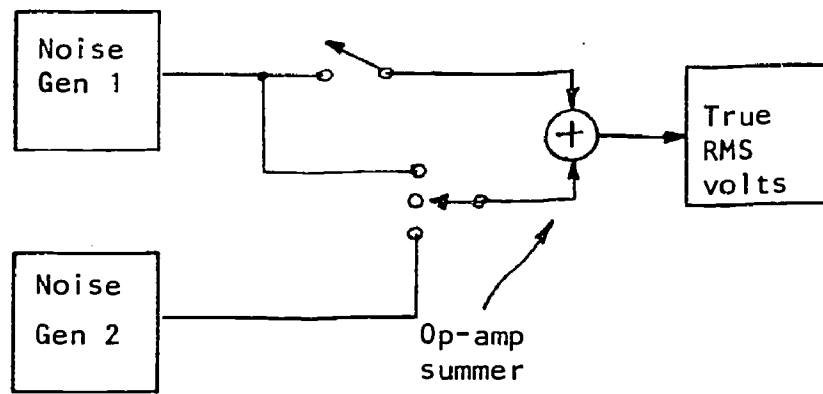
Fig. 5

(d) Use the set-up of Fig. 6 to obtain the cumulative probability distribution function of the noise output of the GR generator. The reading of the output meter (v_o) will be proportional to the distribution function of the noise input. Using the 10 volts full scale and multiplying by 10, its reading will be in percent of the time that $v_n \leq V$, i.e.

$$P(v_n(t) < V)\% = \% \text{ of FS on output meter}$$

Plot the results while taking data (% vs. V) directly on gaussian probability paper (K & E 468000), and on linear coordinate paper. Is the noise close to being gaussian? What is the mean, i.e. the DC offset of the noise generator? Determine the standard deviation from the plot and compare to the rms voltage reading of the noise generator meter.
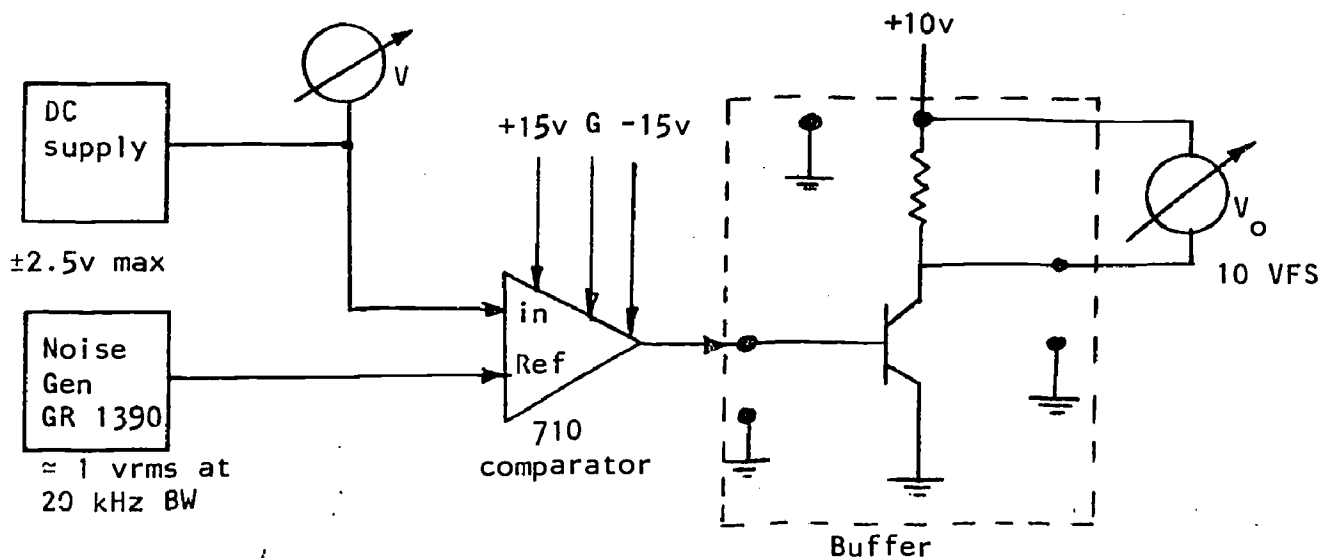

Fig. 6

(e) Optional

By using a storage scope it is possible to directly display the proba-
bility density function.* Fig. 7 shows a suitable set-up. (The reader is
urged to study the reference before attempting this experiment.)
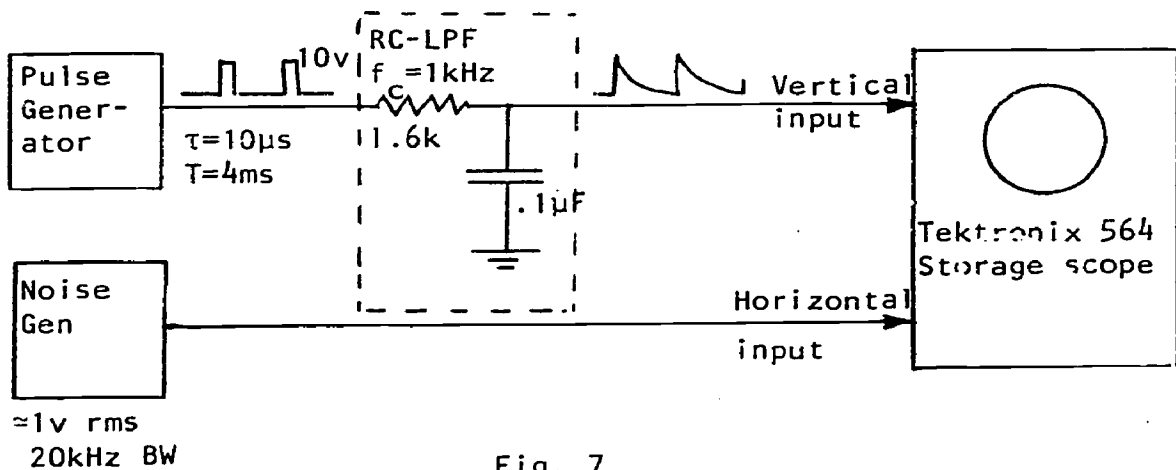

Fig. 7

Using the scope in the non-store mode adjust vertical sensitivity to
get about full scale vertical sweep, and adjust noise generator output to
get approximately full scale horizontal "sweep." Then using the scope in
the store-mode with very low intensity, integrate for a suitable time (5-10
seconds). Several trials will be necessary (with readjustments of the inten-
sity) to obtain a satisfactory result.

Repeat using a sinewave instead of noise. Is the result reasonable?

You may also perform this experiment as part of an optional experiment
towards the end of the semester.

(f) A class demonstration of narrowband noise will be performed by the instructor.

Part II. Noise Figure and Sensitivity of an Amplifier.

Reference: Carlson, Appendix B.

Introduction

Consider the amplifying system in Fig. 8 consisting of a noisy amplifier
with matched (for now) source and load.

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*Algazi, V. R: "Measurements of probabilities and correlations with a storage
   tube," Proc. IEEE, vol. 57, No. 2, (Feb. 1969), pp. 147-159.
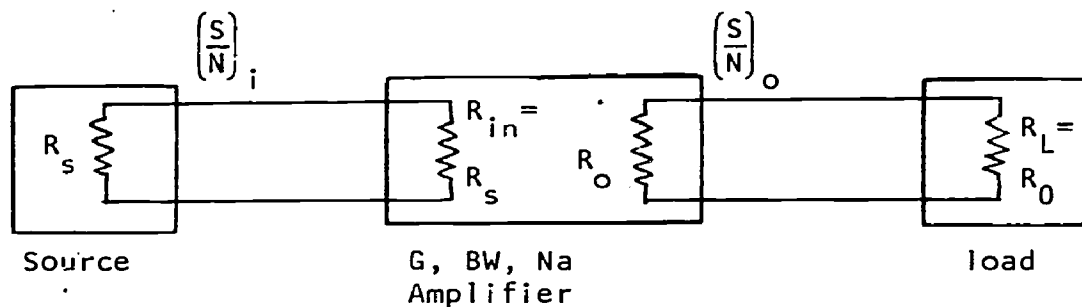
Fig. 8. Amplifying system.

The input to the amplifier consists of the signal plus the noise due to the sour...
The output will consist of an amplified input signal, an amplified input noise,
and the noise generated by the amplifier itself. (Noise generated by the load
may be neglected since it is much smaller than any amplified noise power.)

Let G = available power gain (power gain under matched conditions),

  BW = bandwidth (noise equivalent $\simeq$ 3 dB bandwidth),

  N_a = noise power at output generated by amplifier (excess power noise).

  Let us also assume that the input noise power is white and be represented
by a noise temperature T_i , thus the input noise power to the amplifie; is from
Eqn. (1).

$$N_i = kT_i BW \text{ watts}$$

With a signal input of S_i , the input signal-to-noise power ratio is

$$\left(\frac{S}{N}\right)_i = \frac{S_i}{N_i} = \frac{S_i}{kT_i BW} \tag{3}$$

The output signal power is simply $S_o = S_i G$, and the noise power out is

$$N_o = G N_i + N_a ,$$

thus the output signal-to-noise power ratio is

$$\left(\frac{S}{N}\right)_o = \frac{S_o}{N_o} = \frac{S_i G}{G N_i + N_a} ,$$

or using (3)

$$\left(\frac{S}{N}\right)_o = \frac{1}{1 + (N_a/G kT_i BW)}\left(\frac{S}{N}\right)_i < \left(\frac{S}{N}\right)_i \tag{4}$$

Since $N_a/G\,k\,BW$ depends only on the amplifier, let us define the effective noise temperature (amplifier temperature) by

$$T_e \equiv \frac{N_a}{G\,k\,BW} \quad , \tag{5}$$

(For a noiseless amplifier, $T_e = 0$).

Combining (4) and (5) yields

$$\left(\frac{S}{N}\right)_o = \frac{1}{1 + T_e/T_i}\left(\frac{S}{N}\right)_i \quad . \tag{6}$$

T_e is very useful in describing very low noise amplifiers. For other amplifiers, the <u>noise figure</u> F is more convenient. For the source at <u>room temperature</u> T_o (290°K), F is defined as follows

$$F = \frac{\text{Total noise power in load}}{\text{Load noise power due to source}} = \frac{N_o}{G_a\,N_i} = \frac{N_o}{N_s} = \frac{N_s + N_a}{N_s}$$

$$= 1 + \frac{N_a}{N_s} \geqslant 1 \quad . \tag{7}$$

Show that F may also be written as

$$F = 1 + \frac{T_e}{T_o} = 1 + \frac{N_a}{G\,kT_o\,BW} \quad . \tag{8}$$

Also show that

$$\left(\frac{S}{N}\right)_o = \frac{1}{1 + (F - 1)\,T_o/T_i}\left(\frac{S}{N}\right)_i \quad . \tag{9}$$

If the source is at room temperature, $T_i = T_o$, then (9) reduces to

$$\left(\frac{S}{N}\right)_o = \frac{1}{F}\left(\frac{S}{N}\right)_i \quad . \tag{10}$$

Noise figure is often expressed in decibels:

$$F_{dB} = 10 \log_{10} F \quad . \tag{11}$$

Noise figures of cascaded amplifiers.

Cascading two amplifiers as in Fig. 9, the overall noise figure is given by Friis' formula (see reference page 448)

$$F_{overall} = F_1 + \frac{F_2 - 1}{G_1}$$

(12

where G_1 = power gain of amplifier No. 1

F_1 = noise figure of amplifier No. 1 | operated from same source

F_2 = noise figure of amplifier No. 2 | as the combination
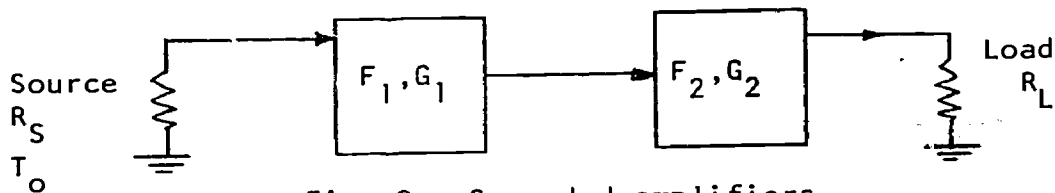
Fig. 9.   Cascaded amplifiers.

As may be seen from (12) if G_1 is sufficiently large then $F_{overall} \simeq F_1$ and the first amplifier determines the noise figure.

Attenuator

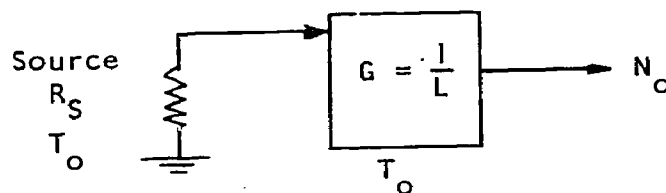Consider a matched resistive attenuator at T_o with loss L, i.e. gain $G = 1/L$, Fig. 10.

Fig. 10.   Matched attenuator.

The noise output is

$$N_o = kT_o BW ,$$

(13

since looking back into the output we see only a thermal resistance.

The noise output can also be written in terms of the effective noise temperat T_e, of the attenuator:

$$N_o = Gk(T_o + T_e)BW .$$

(14

Combining (8), (13), and (14), and using $G = 1/L$, yields

$$F = L \quad , \tag{15}$$

thus inserting an attenuator of loss L in front of an amplifier of noise figure F yields an overall noise figure according to (12) of

$$F_{overall} = L + \frac{F - 1}{1/L} = FL \quad , \qquad \text{or in dB}$$

$$F_{overall} = F_{dB} + L_{dB} \quad . \tag{16}$$

From the above one sees that low-noise systems require a low noise figure preamplifier and a low loss transmission line between the preamplifier and the source (antenna).

Sensitivity of an amplifier.

Sensitivity of an amplifier is often defined as the input signal level necessary for unity signal-to-noise ratio at the output.

Thus with

$$\left(\frac{S}{N}\right)_0 = 1 \quad ,$$

then from (10), the sensitivity is

$$S_i = F \, N_i = F \, kT_o BW \quad . \tag{17}$$

Even with a noiseless amplifier ($F = 1$) the sensitivity is finite since the output still contains noise due to the source. It is interesting to calculate the ultimate sensitivity ($F = 1$). From (17) it is

$$S_i = k \, T_o BW \quad ,$$

and for $T_o = 290^o k$ and $BW = 1$ MHz, this yields

$$S_i = 4 \times 10^{-15} \text{ watts} = 4 \times 10^{-12} \text{ milliwatts} \quad ,$$

or -114 dBm, corresponding to about 0.45 microvolts rms across 50 ohms.

The measurements of noise figure and sensitivity will be discussed in class.

Procedure

a) Measure the noise figure of two Jerrold 406A2 amplifiers cascaded using the HP340B automatic noise figure meter and the HP343A VHF noise source. Fig. 11.
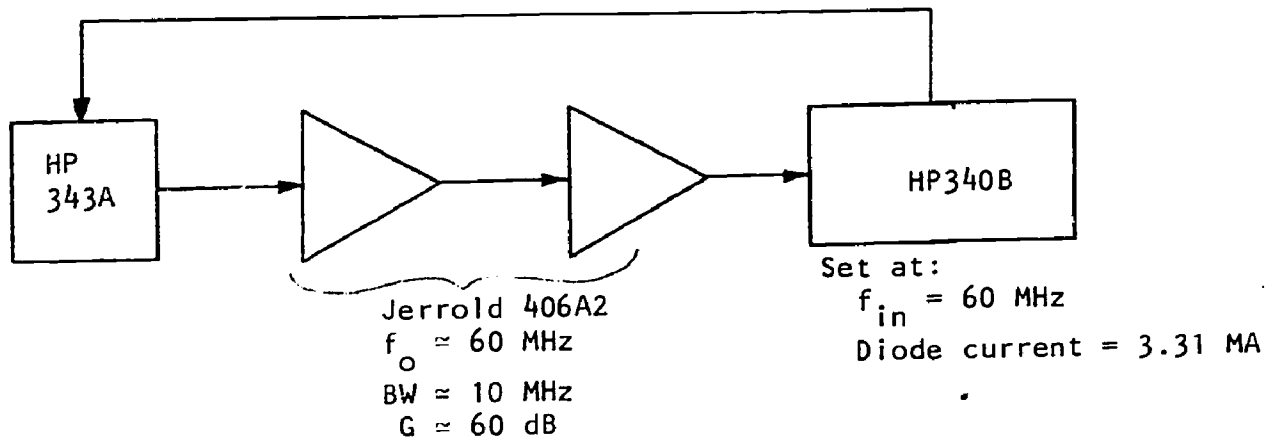
Fig. 11. Measurement of noise figure.

b) Verify eqn. (16) by adding a 3 dB pad to the amplifier chain input and remeasure the noise figure. Repeat for a 6 dB pad.

c) Measure the sensitivity of a baseband amplifier using the set-up of Fig. 12, i.e., determine V_i for $(S/N)_o = 1$.
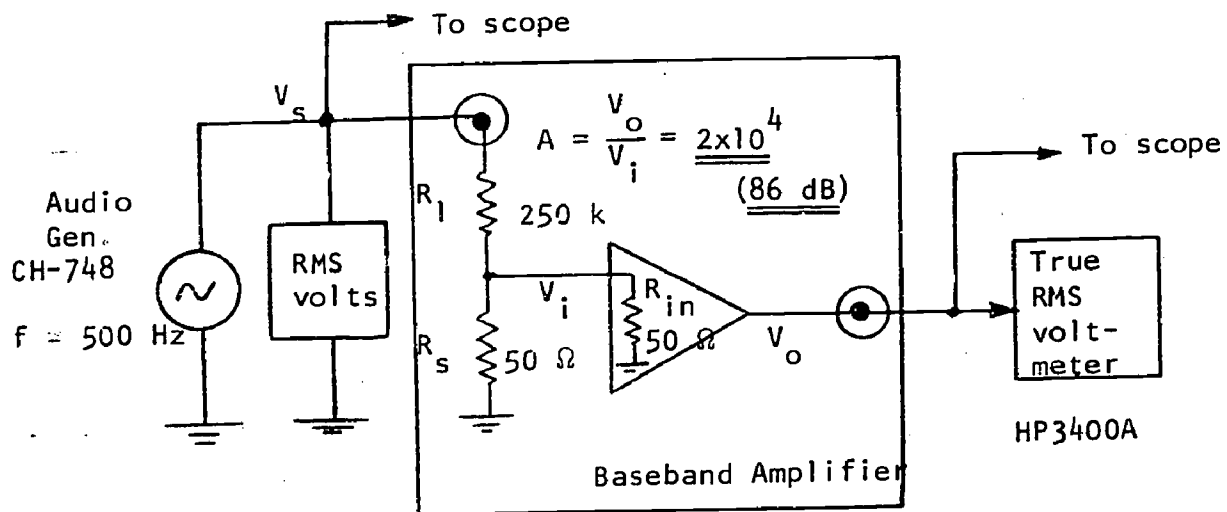

Fig. 12.

The baseband amplifier has a built in input attenuator so as to be able to determine V_i by measuring V_s, and to provide a 50 Ω source impedance for the amplifier (matched condition). From the figure:

$$V_i = V_s \; \frac{R_s || R_{in}}{R_l} = 10^{-4} V_s . \qquad \text{Overall gain} = 2 .$$

d) Determine V_i for a $(S/N) = 100$ (20 dB) .

e) Listen to music reproduction with a signal-to-noise ratio of approximately 1 (0 dB), 10 (10 dB), and 100 (20 dB). Fig. 13.
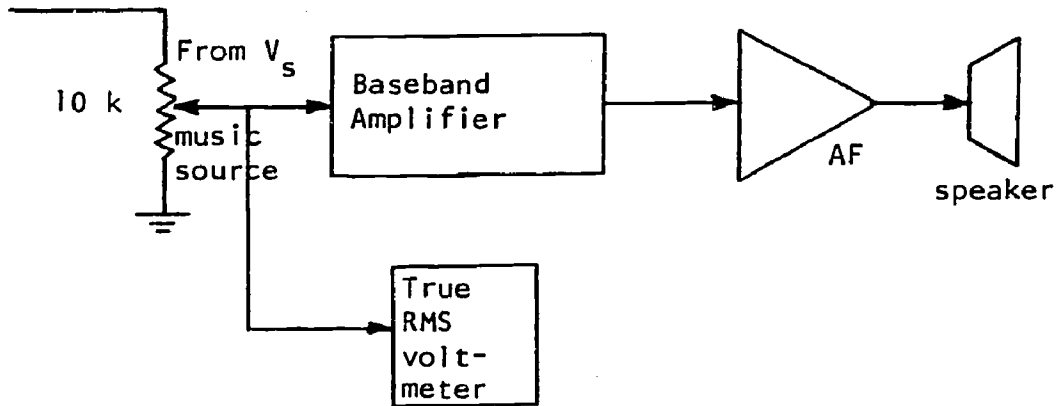

Fig. 13.

CHARACTERISTICS OF RADIO RECEIVERS

Reference: A. Bruce Carlson, Chapter 5.

1. Introduction

Most radio receivers are of the superheterodyne type. A block diagram of a general receiver is given in Fig. 1.
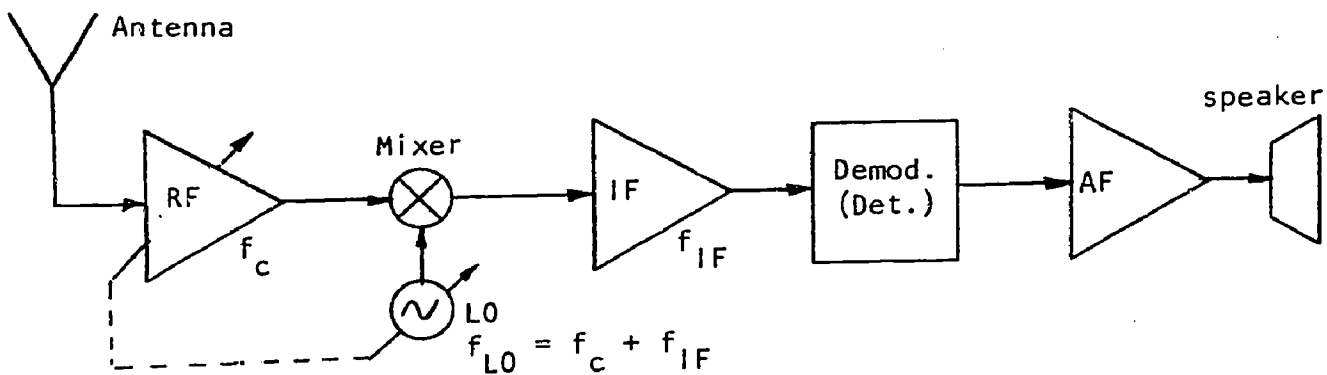

Fig. 1. Superheterodyne receiver

There are three types of amplifiers in a superheterodyne receiver: the radio frequency (RF) amplifier, which is tuned to the desired carrier frequency; the fixed tuned intermediate-frequency (IF) amplifier; and the audio-frequency (AF) or baseband amplifier which provides sufficient level for the speaker or any other output device. The mixer operates as a frequency converter that translates the RF output to the IF band. The detector is a demodulator that recovers the modulating (or baseband) signal and are of a type consistent with the type of modulation used (AM, FM, or PM).

Usually (but not always) the LO frequency is above the signal frequency and the IF amplifier extract the difference frequency $(f_c - f_{LO})$. To better understand the operation of a superheterodyne receiver, consider the frequency domain in Fig. 2.
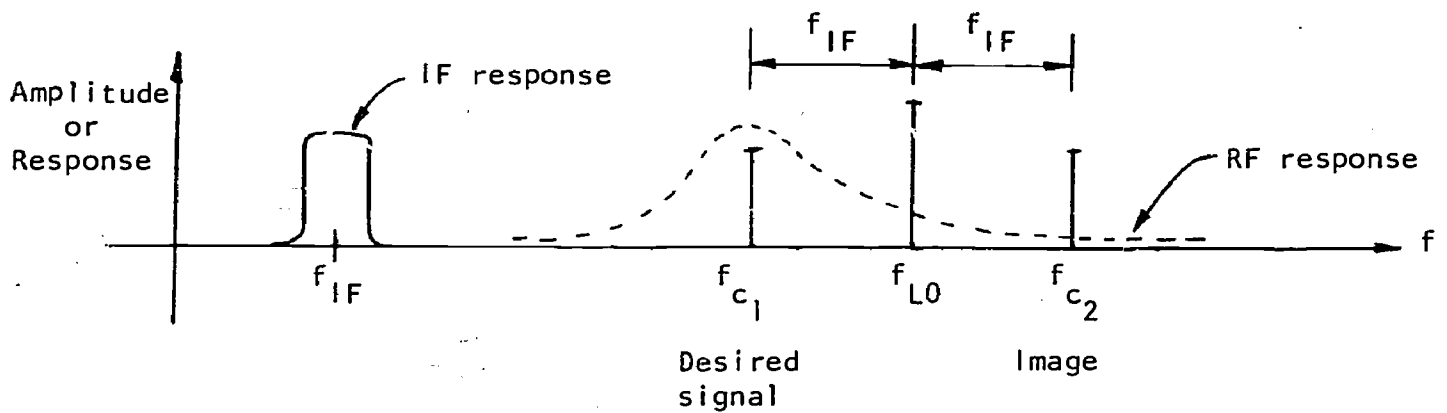
Fig. 2. Superheterodyne Response

With the RF tuned to the carrier frequency f_{c1} and the LO set at $f_{LO} = f_{c1} + f_{IF}$ as indicated, the desired signal will be downconverted to the IF frequency, amplified, demodulated and appear at the output. Note that there is also another signal at carrier frequency $f_{c2} = f_{LO} + f_{IF}$ that will be downconverted ($f_{c2} - f_{LO} = f_{IF}$) to the IF and appear at the output of the receiver. This undesired signal is called the image frequency and produces interference. The purpose of the RF amplifier whose response is shown in Fig. 2, is to attenuate the image as much as possible before it gets to the mixer (and also to provide a low noise figure). As seen from the figure, this requires a large f_{IF}. Thus the RF provides image-channel rejection. Adjacent channel rejection and most of the gain is provided by the IF amplifier whose BW should be no more than the channel-to-channel spacing with sharp skirt selectivity. This requires low f_{IF}.

These conflicting requirements are solved by going to double-conversion in high quality receivers, Fig. 3.
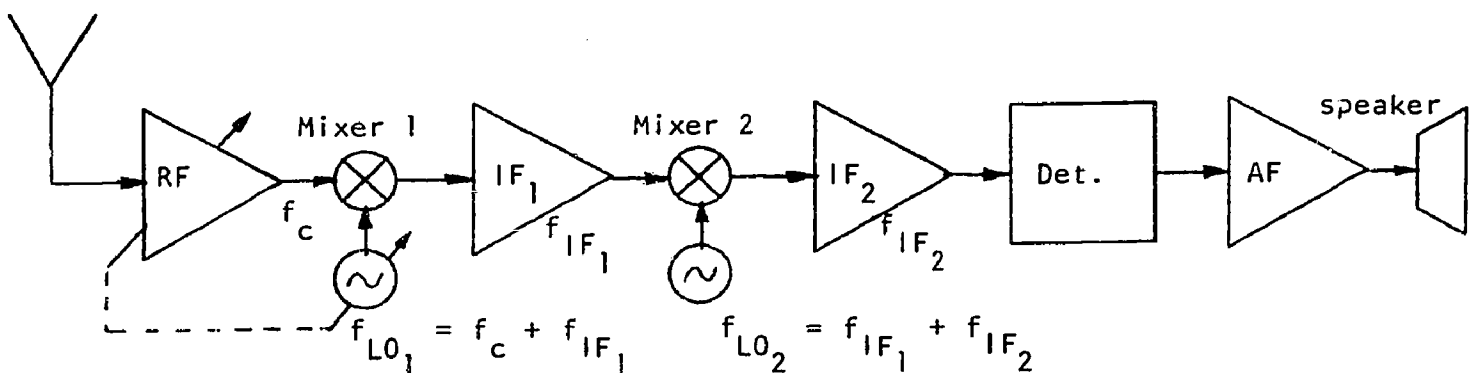

Fig. 3. Double-conversion superheterodyne receiver

In the double-conversion receiver the first IF is tuned to a large f_{IF_1}, so as to get good image rejection, while the second IF has a much smaller f_{IF_2} to provide good channel selectivity.

Besides the image response, receivers suffer other undesired responses called spurious responses. Some spurious responses are due to non-linearities, imperfections of the various devices that make up a receiver, and the fact that the LO signal contains harmonic components. In general, one would expect spuriouses at (or near) $\frac{1}{2}f_{IF}$, f_{IF}, and $|nf_{LO} \pm mf_{IF}|$ with n and m integral.

2. The AM receiver

In the AM receiver all the amplifiers operate in their linear range in order to preserve amplitude information. The detector is a simple envelope detector for AM-DSB or a synchronous detector for AM-DSB/SC and AM-SSB/SC. An automatic volume control (AVC) is also added.

Equipment available

Unknown brand solid state x-tal controlled, single-conversion AM receiver for signal frequency about 40 MHz. It consists of an RF amplifier, mixer, IF amplifier, envelope detector, and AF amplifier. The LO is above the signal frequency. The IF frequency is 455 kHz. (2 units)

The Regency TMR-1A is a x-tal controlled double-conversion AM receiver designed for the aircraft band from 118-136 MHz. It consists of an RF amplifier, 2 mixers, 2 LO's, 2 IF amplifiers, a regular envelope detector and an audio amplifier. The first IF frequency is 10.5 MHz and the second is 455 kHz. The first LO is above the signal frequency. See instruction manual for further details. (3 units)

Procedure

(a) Sensitivity

Measure the sensitivity in dBm and microvolts using 30% modulation at 400 Hz. Sensitivity is defined as the input signal necessary to produce an audio S + N/N ratio at the output of 10 dB. Fig. 4.
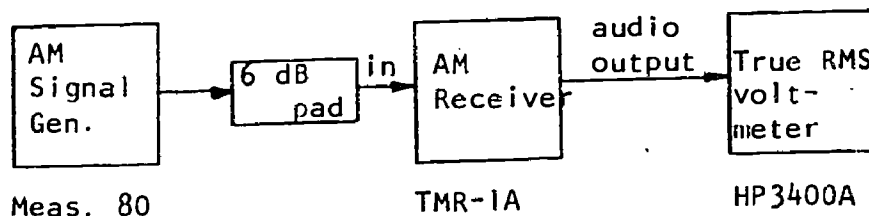

Fig. 4. Receiver Measurements

(b) Image Rejection

      Measure the image rejection in dB with 30% modulation: set the signal generator at the image frequency and measure the image sensitivity (as in (a)). The difference between the sensitivities (signal frequency and image frequency) in dBm is the image rejection in dB.

(c) Spurious Responses

      Try to find some spurious responses and measure their rejections. It will be necessary to use strong input signals.

3. The FM receiver

In an FM receiver the detector is an FM demodulator of the form described in Exp. No. 6. The detector should ideally only respond to variations of the frequency of the input signal; however, most FM detectors also respond to amplitude variations (an exception is the PLL). In order to eliminate its response to amplitude variations which would produce added output noise, a limiter is inserted ahead of the detector and the IF amplifier gain increased to the point when the receiver limits on just noise. Thus in an FM receiver the IF/limiter combination operates in a non-linear region as opposed to the AM receiver, thus AVC is not desirable or used. On the other hand automatic frequency control (AFC) is sometimes used to keep the receiver tuned to the desired signal.

In the absence of an input signal the input to the detector is amplitude limited noise whose phase is random. Since the time rate of change of phase corresponds to frequency the detector output is also noise. In Fig. 5 is shown a noise phasor whose endpoint may be anywhere within the circle of amplitude limits, and thus it will produce quite a bit of noise output of the detector.
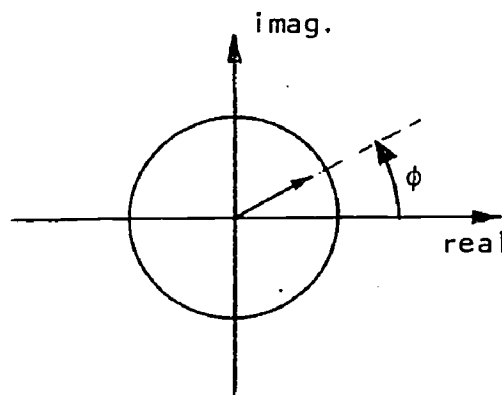

Fig. 5. Amplitude limited noise phasor
Phase variation = 2π radians

With just a carrier (no modulation) received the situation changes radically. This case is shown in Fig. 6.
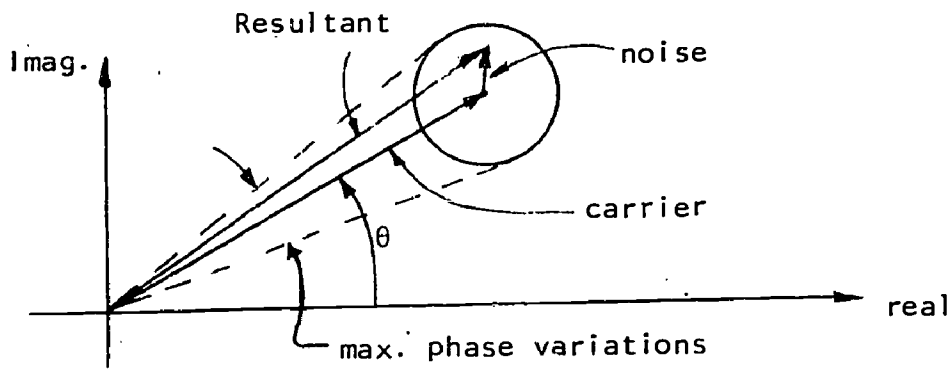

Fig. 6. Carrier plus noise

If the carrier is considerably larger than the noise, the carrier essentially controls the phase of the resulting phasor and the random phase variations are much reduced and hence the detector noise is reduced. This is called quieting. The onset of quieting is quite sharp and is called the FM threshold. Above the threshold the output signal-to-noise ratio is very good, below threshold it is so poor as to be useless for communication. Quieting is often used for FM receiver sensitivity measurements.

The threshold may be extended by going to frequency feedback and the use of PLL*.

As pointed out in Exp. No. 6, the audio signal power increases with an increase of Δf without changing the transmitter power (as long as the FM signal is above the threshold). Thus, increasing Δf increases the audio output signal-to-noise ratio. The increased Δf results in an increased transmission BW, hence FM belongs to a class of modulation systems designated wideband noise reduction systems.

Equipment

Model PF-175 is an VHF-FM tunable single-conversion receiver with an IF frequency of 10.7 MHz. The LO is above the signal frequency. Use this receiver at about 32 MHz signal frequency. (1 unit)

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*H. Taub and D. L. Schilling: "Principles of Communication Systems," Chapter 10, McGraw-Hill, 1971.

Model Dura-Scan-4 are VHF-FM x-tal controlled double-conversion receivers. The first IF frequency is 10.7 MHz and the second is 455 kHz. The first LO is below the signal frequency. (3 units)

For more details on these receivers see the instruction manuals.

Procedure

Using the block diagram of Fig. 7 do:

(a) Measure the sensitivity. Sensitivity is defined as the unmodulated carrier input level (in dBm or microvolts) necessary to reduce the audio output noise level by 20 dB from its value with no carrier input.

(b) Measure the image rejection in dB. Set the signal generator at the image frequency and measure the image sensitivity as in (a). The difference between this value in dBm and that found in (a) in dBm is the image rejection in dB.

(c) Measure the discriminator (detector) demodulation sensitivity (slope). Using an input level about 20 dB higher than that found in (a) modulate with a 1 kHz tone at a deviation of $\Delta f = 16$ kHz for PF-175 or $\Delta f = 2-3$ kHz for Dura-Scan-4 and measure the discriminator output voltage. Compute the demodulation sensitivi from these data.
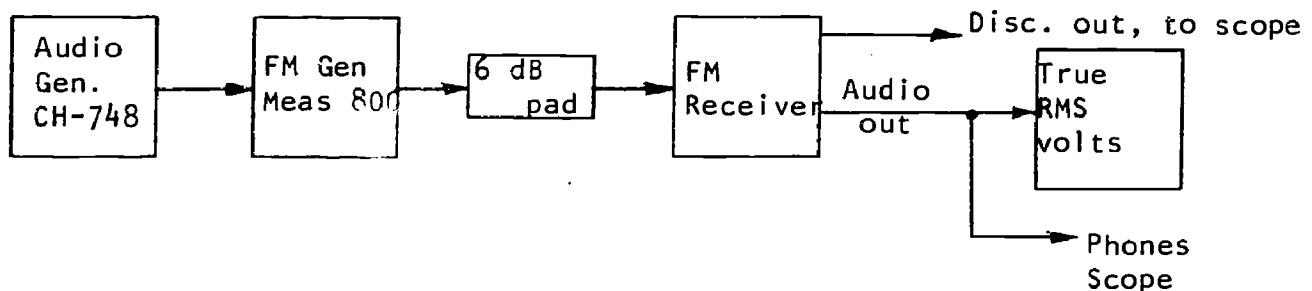
Fig. 7.   FM receiver measurements

If time allows do the following:

(d) Measure the audio frequency response using a modulated RF signal. Plot the response in dB vs. log f (semilog paper).

(e) Check for spurious responses.

Using the block diagram of Fig. 8, do:

(f) Using a sweep frequency technique display the discriminator S-curve on a scope. Calibrate the scope display and photograph. Compute the demodulation sensitivity. Fig. 8.
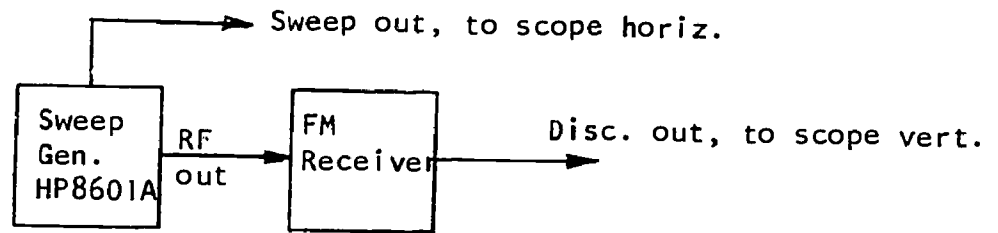
Fig. 8. S-curve display

(g) You may also test your own FM broadcast tuner if you like.

EXPERIMENT NO. 9

DIGITAL MODULATION

Reference: Chapter 9, Carlson.

Introduction

The only difference between digital modulation and analog modulation is that in digital modulation the baseband or modulating signal is a digital waveform instead of an analog waveform. The digital information is impressed upon a carrier in the same way as in analog modulation, resulting in digitally modulated AM, FM, or PM. The digitally modulated techniques are termed amplitude-shift keying (ASK), frequency-shift keying (FSK), and phase-shift keying (PSK) to distinguish these from regular analog AM, FM, and PM. Generally, however, the modulation methods differ little, if any.

Given a digital message, the simplest modulation technique is ASK where the carrier amplitude is switched between two or more values, usually on and off pulses for binary signals. Fig. 1 depicts a typical binary unipolar signal and the resulting ASK, FSK, and PSK waveforms corresponding to rectangular pulse modulation of an AM, FM, or PM modulator, respectively.

These simple forms are not practical since they result in excessive transmission BW owing to the use of a rectangular data signal. The transmission BW is reduced by shaping the pulses (LPF) before modulation. Furthermore, it is better to go to a bipolar baseband signal so as to eliminate the DC offset of a unipolar signal. Pulse shaping is done not only to reduce the transmission BW but also to reduce intersymbol interference which may occur due to imperfect channels. (See Nyquist pulse shaping, page 381, Carlson.)
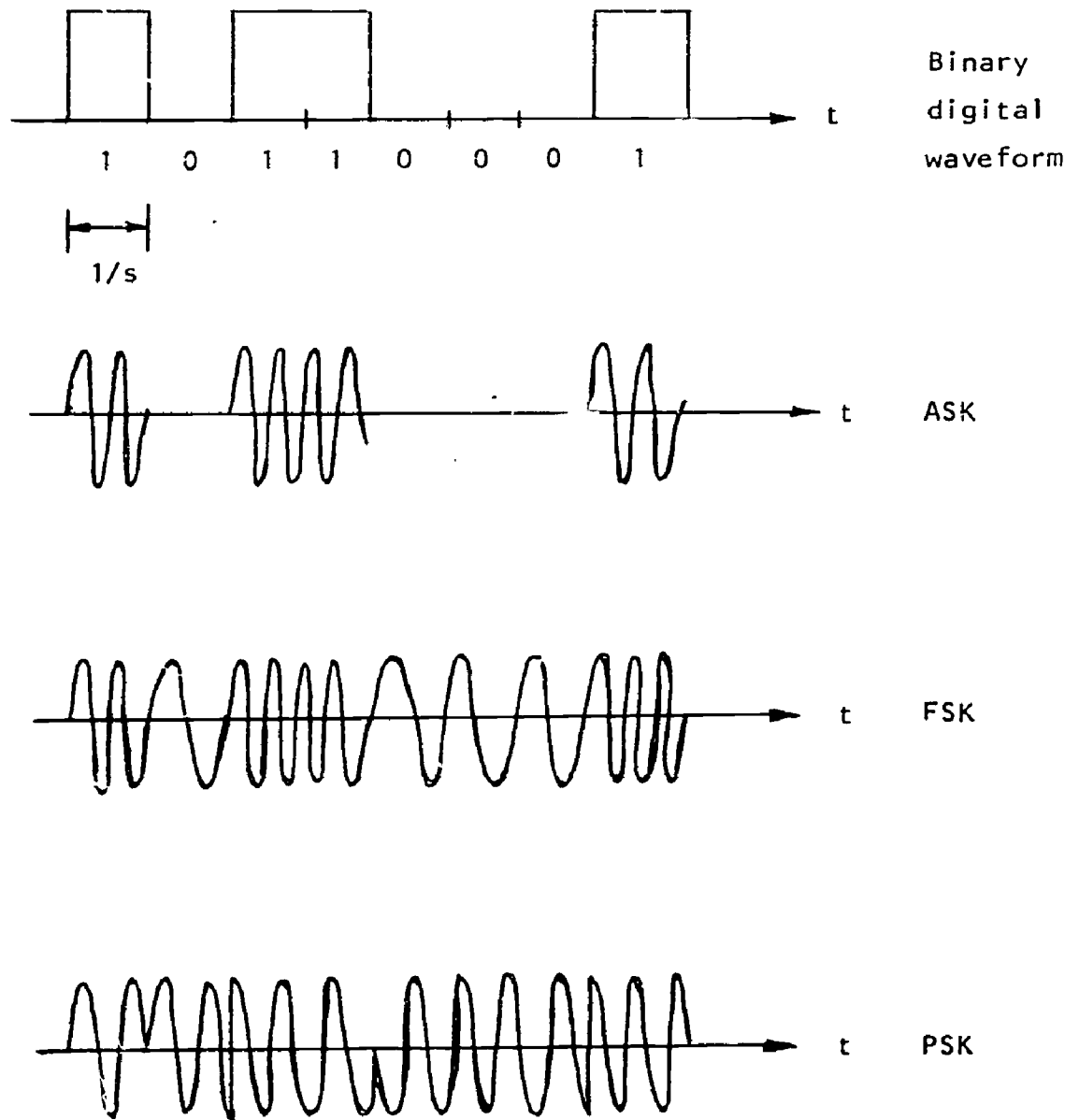
Fig. 1. Unipolar binary baseband signal for the message 10110001
and the resulting digitally modulated signals.
s = signalling rate

Fig. 2 shows a typical rectangular bipolar digital signal and the resulting filtered or shaped waveform.

Unshaped
signal

1   0   1   ?   0   0   0   1
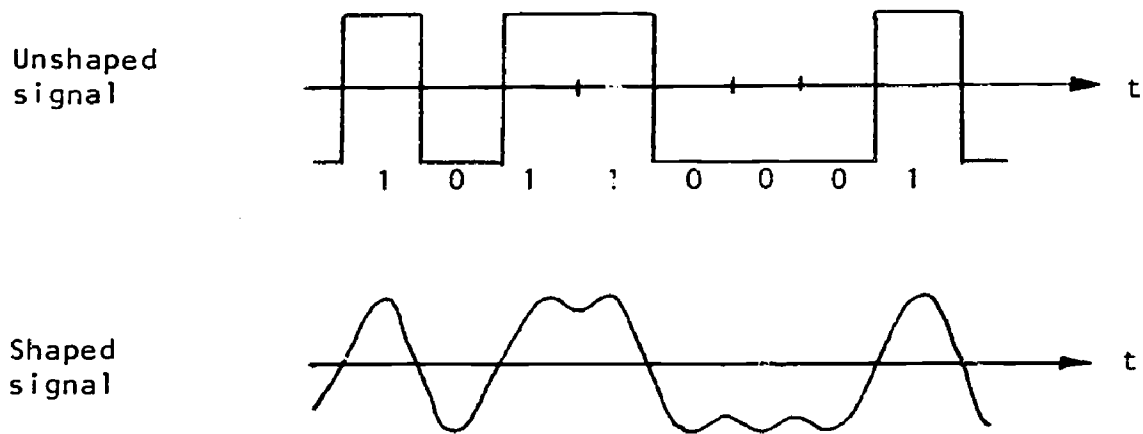
Shaped
signal

Fig. 2.   Typical bipolar baseband signals

Using the shaped signal of Fig. 2 in an AM-DSB/SC modulator results in the signal shown in Fig. 3.

Fig. 3.   Binary DSB or PRK signal

This signal has phase reversals whenever the modulating signal changes sign; thus it is a PSK signal with 180° phase changes. For this reason, it is called phase-reversal-keying (PRK). Notice that it also has amplitude changes. Demodulation requires a synchronous (or coherent) demodulator just as in the case of AM-DSB/SC.

Problem: Sketch the resulting FSK signal using the shaped baseband signal
of Fig. 2.

It can be shown (see Carlson) that bipolar baseband coherent PSK is the
best system of the three types described with respect to resulting in the lowest
error probabilities in the presence of gaussian noise for a given signal-to-
noise ratio.   The problem with coherent PSK is of course that it requires precise
knowledge of the carrier phase and frequency, thus involving more sophisticated
hardware.

A technique known as differentially coherent PSK (DPSK) has been devised
to get around the synchronization problems of coherent PSK.   The reader is
referred to the reference for details and discussions of the relative merits
of this system.

Procedure

1.   Demonstrate digital modulation by generating the waveforms of Fig. 1:
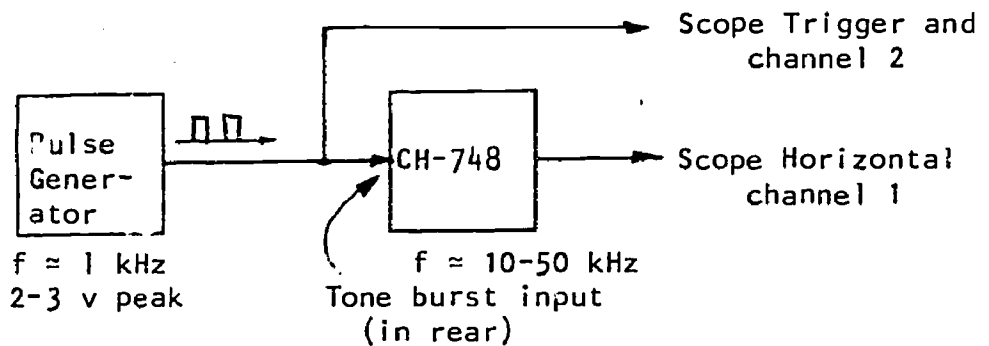a.   Generate ASK using the set-up of Fig. 4.

Fig. 4.   ASK generation
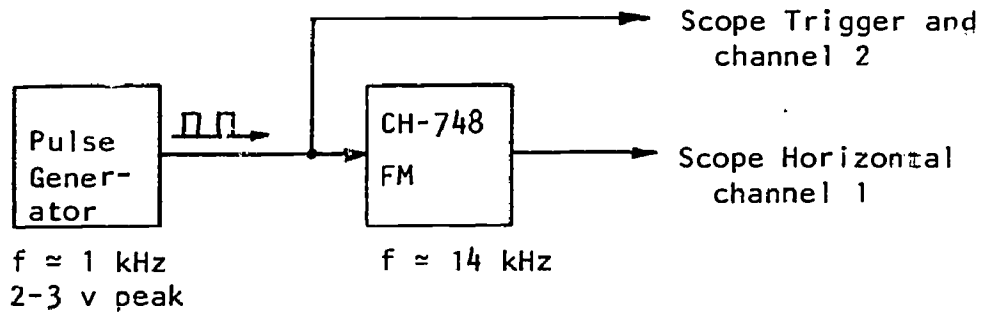
b.   Generate FSK using the set-up of Fig. 5.

Fig. 5. FSK generation

Adjust the frequency of the CH-748 AM/FM generator to obtain an approximately stationary pattern on the oscilloscope.

c. Generate PSK using the set-up of Fig. 6.
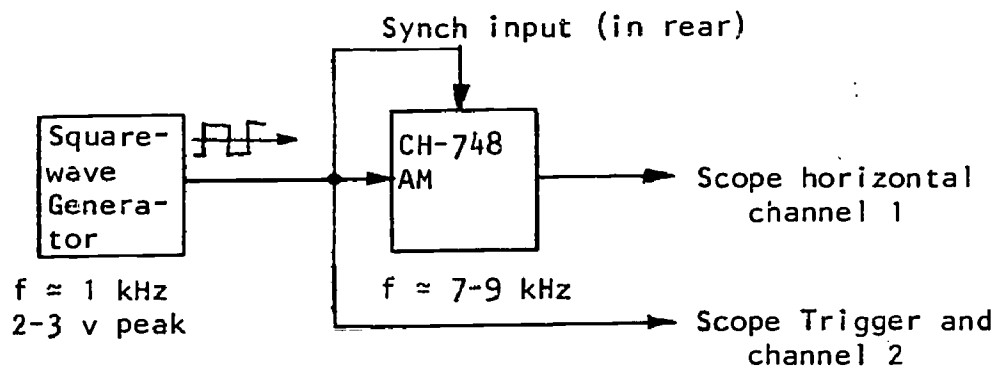
Fig. 6. PSK generation

Adjust the frequency of the CH-748 carefully to obtain lock at a harmonic of the square wave. When locking has been obtained a stationary waveform similar to that shown in Fig. 1 will be displayed on the scope. Notice the phase reversals at the square wave transition times.

2. Demonstrate synchronous demodulation of the PSK signal generated in 1c. Fig. 7.

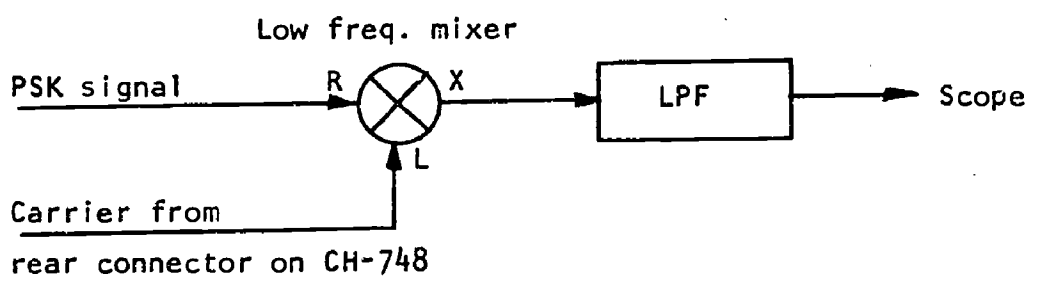The unmodulated carrier is available from a rear connector on the CH-748.

Low freq. mixer

PSK signal   R ⊗ X   [LPF]   → Scope

           L

Carrier from

rear connector on CH-748

Fig. 7.   Coherent demodulation of PSK

EXPERIMENT NO. 10

MICROWAVE HORN OR APERTURE-ANTENNA PATTERNS

Objective: To make an experimental study of the far-zone pattern characteristics of particular horn configurations at microwave frequencies, and to compare the experimental results with mathematical expressions for the patterns.

Introduction

An "antenna" may be considered as a device used to couple an electromagnetic guided wave into space.* The space into which an antenna couples cannot always be considered as "free (unbounded) space," however. Many practical situations involving antenna applications must take into account the presence of the earth and its irregularities (buildings and mountains), as well as the possible effects of the free electrons and gas ions in the rarefied atmosphere at high altitudes (the ionosphere). Such irregularities may play an important role in determining the amount of energy coupled into a receiving antenna from a remote transmitting antenna.

At radio frequencies the effect of the earth or sea, including the spherical shape of the earth, as well as the ionosphere, are important in energy transmission over an open path. At microwave frequencies (about 1000 MHz and up), the problem of transmission over essentially line-of-sight paths is of prime importance.** (Note Fig. 1.)

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*In a special sense, however, an "antenna" need not couple a wave into "space" as such; for example, the short output wire of the 2K25 klystron, which couples the TEM wave of the output coaxial line of that tube into a rectangular waveguide, can certainly be regarded as an antenna. Instead of coupling into space, it couples into a bounded region.

**From an information theory viewpoint, the remarkable bandwidth properties (and hence channel-handling properties) of microwaves makes them practical for certain communication or data-transmission purposes, even though many relay stations may be necessitated by the line-of-sight transmission-distance limitation. At the same time, the very high frequencies also make the use of very high gain, narrow-beam antennas practical, and so relatively low-power transmitters can be utilized.
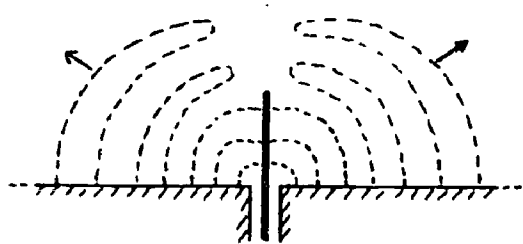
(a) Vertical monopole antenna above earth, showing how earth behaves as a "lossy waveguide."
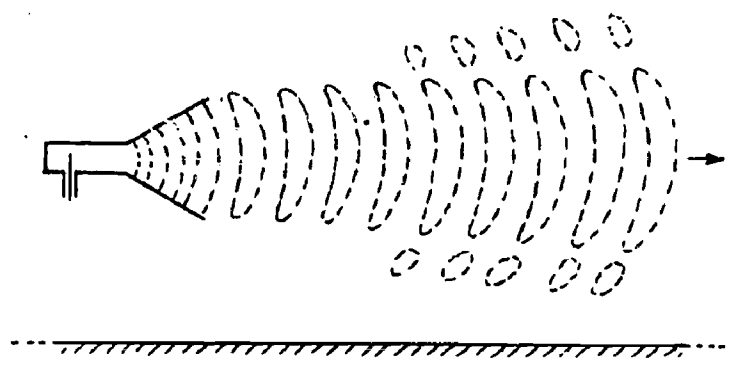
(b) Horn antenna above earth, depicting the radiated field behavior which may be largely independent of the presence of the earth.

Fig. 1. Examples of antenna devices for which the earth plays (a) important, and (b) lesser roles.

Antenna configurations useful for microwave applications are manifold indeed. They take innumerable forms, varying from slot arrays cut into waveguide walls to large parabolic metallic reflectors or dielectric lenses fed by waveguide horns. Our present attention is focused on the far-zone field theory of the waveguide horn antenna, since the theory of this device is closely related to the performance of many other "aperture" antennas. Furthermore, horns both large and small have extensive microwave applications and are comparatively easy to build.

The problem of calculating or predicting the electromagnetic field in region exterior to a horn may be shown to depend upon the knowledge of the tangential electric and magnetic fields over any closed surface enclosing the electromagnetic sources (see Ref. 1). Early applications of this (or related) principle to the electromagnetic horn include those of Barrow and Greene (Ref. 2) and of Barrow and Chu (Ref. 3), as well as an equivalent approach described by Schelkunoff (Ref. 5), and an account of both experimental and calculated results pertaining to horns as well as other microwave antennas is given in the book by Silver (Ref. 4). Refer to Fig. 2, and note that if the closed surface S encloses the rectangular horn in the manner shown, then tangential $\bar{E}$ and $\bar{H}$ fields on S contributing to the field at any farzone point P are substantially those which occur in the horn
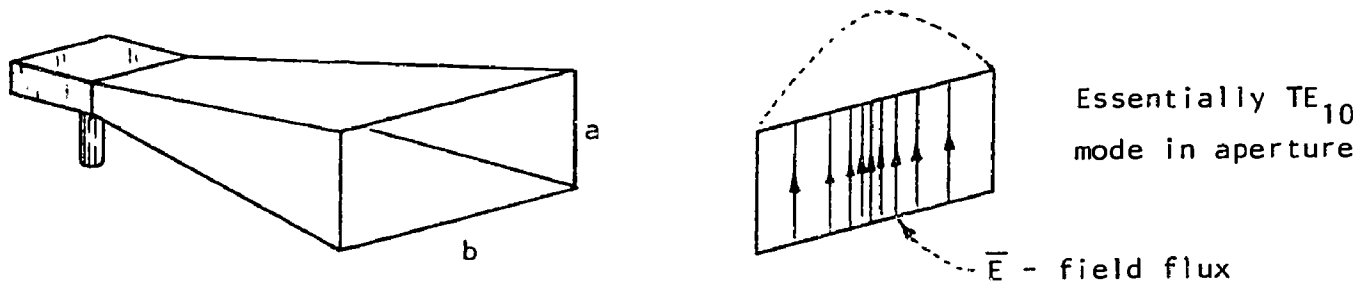
(a) A microwave horn.
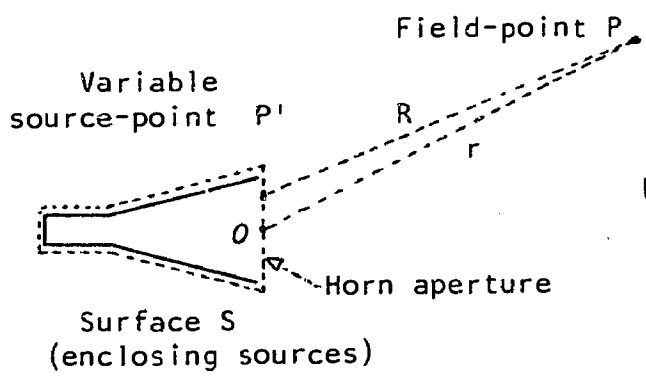
(b) Showing surface S constructed around horn, preparatory to integrating for $\overline{E}$, $\overline{H}$ field at P resulting from tangential $\overline{E}$, $\overline{H}$ on S.
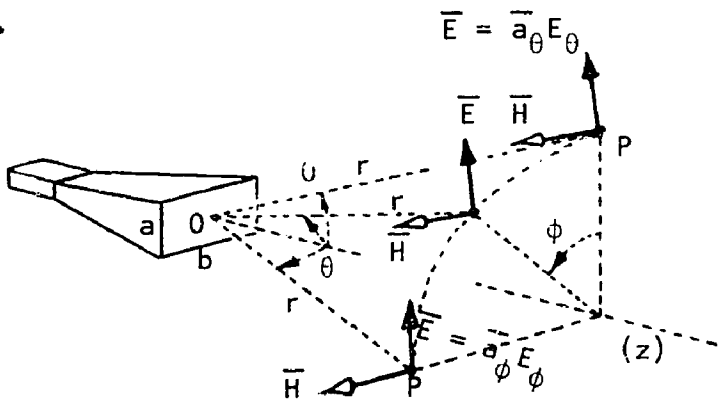
(c) Geometry of horn, and spherical coordinate system for locating "far-zone" field point P, distance R away. Note principal (E and H) planes.

Fig. 2. Details of pattern development for
a Rectangular waveguide horn.

aperture; on the exterior metal horn surfaces, the tangential $\overline{E}$ field is zero and tangential $\overline{H}$, associated with induced currents on the horn exterior, is considered negligible. Then, supposing the aperture fields to consist primarily of a TE_{10} mode, one can integrate for the fields at farzone point P, by methods given in the References, to obtain the following in the two principal planes of Figure 2(c).

(1) In the E-plane of Fig. 2(c) (where $\phi = 0^{o}$):

$$\hat{E}_\theta = \frac{j\beta_o ab\hat{E}_m}{8r} (1 + \frac{\beta_{10}}{\beta_o} \cos \theta) \frac{\sin A}{A} e^{-j\beta_o r} \;,\; -\frac{\pi}{2} < \theta < \frac{\pi}{2} \;. \quad (1)$$

(2) In the H-plane of Fig. 2(c) (where $\phi = 90^\circ$):

$$E_\phi = \frac{j\beta_o ab\hat{K}}{8r} \left(1 + \frac{\beta_{10}}{\beta_o} \cos\theta\right) \frac{\cos B}{B^2 - (\frac{\pi}{2})^2} e^{-j\beta_o r} ,$$

$$-\frac{\pi}{2} < \theta < \frac{\pi}{2} . \tag{2}$$

Here, $\beta_o = \frac{2\pi}{\lambda}$, and β_{10} = horn-waveguide phase constant in the aperture for

TE_{10} mode;

a,b = horn height and width as in Figure 2;

$\hat{E}_m$ = E-field amplitude in horn aperture;

r = distance to field point from horn-aperture center;

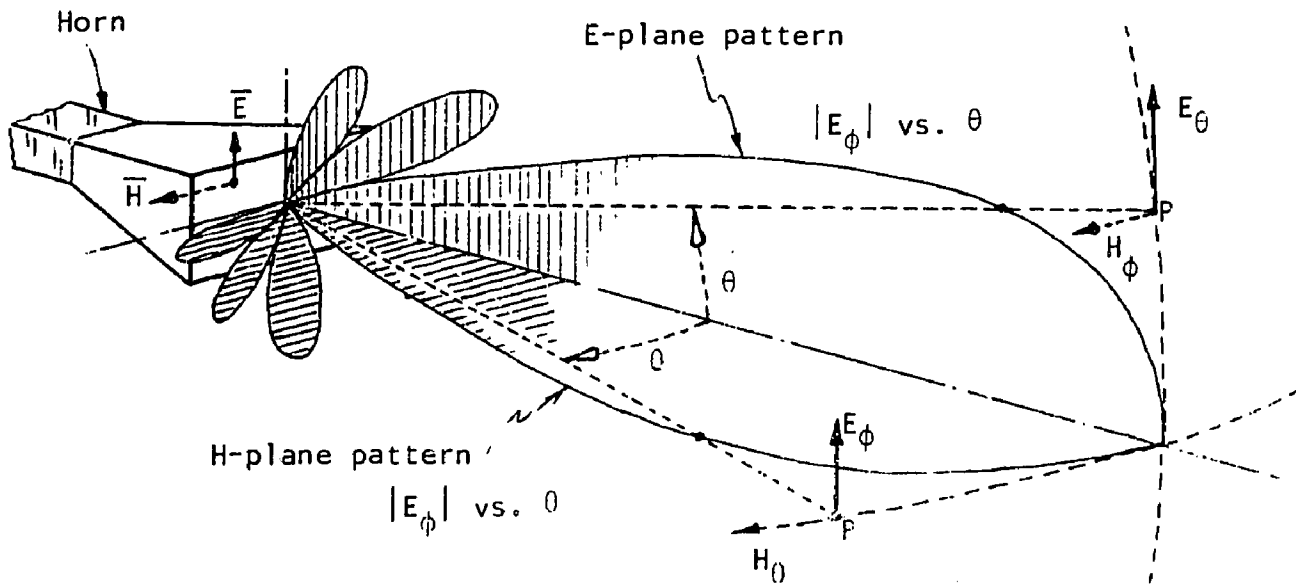$A = \frac{1}{2}\beta_o a \sin\theta = \frac{\pi a}{\lambda}\sin\theta$;

$B = \frac{1}{2}\beta_o b \sin\theta = \frac{\pi b}{\lambda}\sin\theta$.

Results (1) and (2) are good only for the far-zone region; i.e., where r is
very large compared to horn dimensions a, b . The $\overline{H}$ intensities corresponding
to E_θ and E_ϕ of Eq. (1) and (2) are transverse to $\overline{E}$ as shown in Fig. 2(c),
and related to $\overline{E}$ by the intrinsic impedance $(\mu_o/\epsilon_o)^{\frac{1}{2}}$ for the homogeneous medium
in which point P is located. Note that E falls off as r^{-1} in the far zone;
so the power density related to the Poynting vector falls off as r^{-2} .

Functions (1) and (2) take on a simpler significance in the light of their
"antenna pattern" interpretation. Holding R fixed, the $\overline{E}$ field magnitudes
in the principal planes are proportional to the following so-called "field pattern"
angular functions:

(1) E-plane: $|\hat{E}_\theta| \propto \left(1 + \frac{\beta_{10}}{\beta_o} \cos\theta\right) \left|\frac{\sin A}{A}\right|$, $-\frac{\pi}{2} < \theta < \frac{\pi}{2}$. (1a)

(2) H-plane: $|\hat{E}_\phi| \propto \left(1 + \frac{\beta_{10}}{\beta_o} \cos\theta\right) \left|\frac{\cos B}{B^2 - (\frac{\pi}{2})^2}\right|$, $-\frac{\pi}{2} < \theta < \frac{\pi}{2}$. (2a)

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(a) Diagrammatic representation of polar antenna patterns, showing relation of received signal intensity $|E|$ to θ-position, in two principal planes.
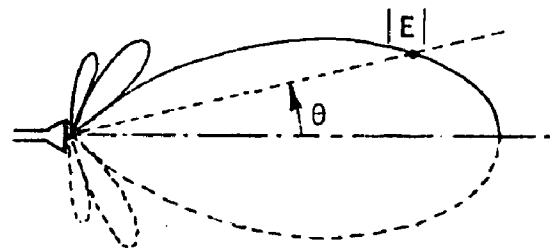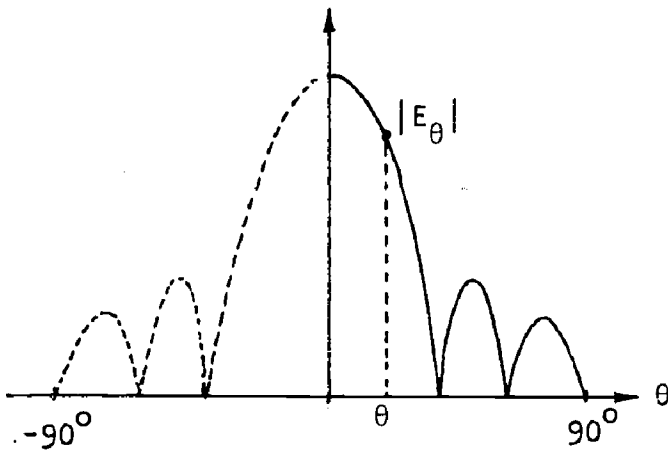
(b) Rectangular-coordinate pattern plot.   (c) Polar-coordinate pattern plot.

Fig. 3. Methods of antenna pattern representation.

These pattern functions (1a) and (2a) are shown sketched for a typical horn in Fig. 3. A little study of the function shows that the _larger_ the aperture dimension a , the _narrower_ will be the width of the main beam. For example, if a = 5λ , then the main-beam-width between half-power points (or 0.707 E_{max}-points) becomes B.W. = 10° approximately; again, if a = 10λ , then B.W. = 5°, and so on. (This may be termed a "five-and-ten" or "Woolworth" rule, which holds for _rectangular_ apertures having constant aperture fields in the direction aligned with that principal plane.)

Fig. 3 shows how typical antenna patterns may be sketched in either polar or rectangular coordinates. It is sometimes desirable to plot the $|E|$ axis in relative dB units, so as to give amplified side-lobe information, which may be quite important for some applications. In Fig. 2(b), note the sketch of a typical experimental far-zone field pattern, seen to deviate from the calculated pattern expecially in the side-lobe region, where only dips are shown instead of distinct nulls. These deviations are the result of the approximations made in assuming the aperture fields which led to Equations (1) and (2); if higher modes* in the aperture were properly accounted for (in terms of precisely satisfying boundary conditions inside and outside the horn), then better correlation between the experimental and calculated patterns would be expected.

Because of the relatively narrow-beam property of rectangular waveguide horns, pattern measurements can be made on a horn by rotating it in the principal plane of interest, and measuring the field produced at a fixed distance r as θ is changed.**

Laboratory Procedure

Connect the apparatus as shown in Figure 4 of the experiment. The klystron should be modulated with a 1 kHz square wave. The polar angle on the plotting paper should read zero degrees when the test horn is directly aligned with the

---

*Such modes, in fact, must be identified with the _tapered_ boundaries of the horn, and not a straight waveguide. See Reference 4 in this connection.

**In practice, the horn whose pattern is to be measured may be used as the "receiving" horn. The transmitter, a distance r away, is considered to be capable of producing an almost uniform, plane-wave field at the receiving-horn position. That the antenna patterns on "receive" and on "transmit" are the same follows from the theorem of reciprocity for antennas.

transmitting horn (maximum deflection of the amplifier). Consult the instructor
concerning the adjustment of the automatic pattern plotting system.

1. Horn Pattern Measurements. Obtain polar plots of the relative electric
    intensity versus θ for the two principal planes of rectangular test horns
    supplied (for ϕ equal to zero and ϕ equal to 90°: E-plane and H-plane
    patterns). It may be necessary to use two twisted sections of waveguide to
    secure correct polarizations. (Let R = _____ meters.) Note that several
    sets of data may be recorded on a single graph. Polar graph paper ($8\frac{1}{2}$ x 11)
    may be secured at the bookstore, if desired, though plain paper is satisfactory.

2. Patterns of Other Antennas. Obtain pattern data for other antennas which
    the instructor may supply. (Circular parabolic dish antennas of different
    sizes are possibilities.) Record these patterns on auxiliary polar graphs.

3. Cross-Polarization Effects. Turn the test horn so it is vertically polarized
    and the transmitter is horizontally polarized and observe how much signal
    is received when compared with the proper polarizations of the antennas.
    Comment on the observations.

4. Measure the dimensions of the test horn, and check the operating frequency
    of the transmitter. Express the horn (or aperture) dimensions in terms of
    free-space wavelength λ_o (e.g., a/λ_o , etc.).

Summary

Compare the beam widths with those calculated by use of the "5-and-10" rule,
where applicable. Comment on the general effect of aperture size on beamwidth
for rectangular horns. Are similar rules applicable to circular apertures?
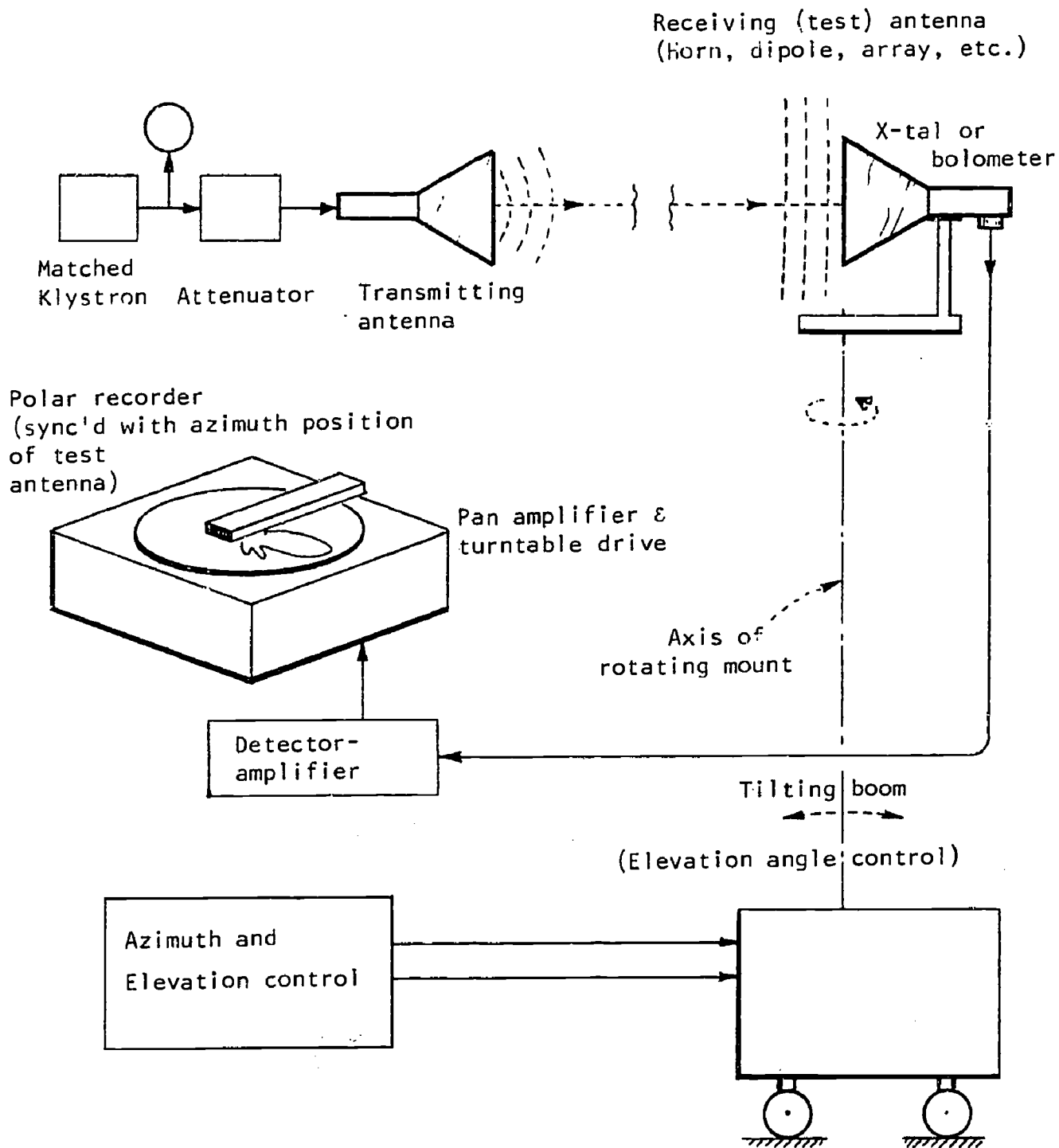What are the advantages of a very narrow "pencil" beam in microwave communication?

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Fig. 4. Antenna pattern measuring system.

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EXPERIMENT NO. 11

RECEPTION OF CLOUD-COVER PICTURES FROM WEATHER SATELLITES

AN EXAMPLE OF A SIMPLE COMPLETE ANALOG COMMUNICATION SYSTEM
USING SUBCARRIER MODULATION

Historical background

The weather satellite era began in 1963 with the launch and flight-testing
of the first automatic picture transmit (APT) TV camera subsystem on TIROS VIII.
From 1963 to 1968 several satellites were launched and designated ESSA-1,2,3,...8
(Environmental Science Services Administration). Some of these satellites failed
either during launching or in orb't; however, most of these provided a wealth of
cloud cover pictures to more than 800 ground stations throughout the world. The
last satellite in this series (ESSA-8) was deactivated due to failure of the
camera shutter mechanism on March 12, 1976, after more than seven years continuous
operation. The replacement satellites for the ESSA series are the NOAA satellites
(National Oceanic and Atmospheric Administration. As of this writing there are
two active NOAA satellites (NOAA-3 and 4). The NOAA (as well as the ESSA) satel-
lites are in polar orbits with a period of approximately 115 minutes. In addition,
there are a number of geostationary (24 hour orbital period) satellites: ATS-1,
ATS-3, SMS-1, and SMS-2. The ATS satellites are basically used for relaying
weather charts and cloud-cover pictures, while the SMS satellites also have on-
board cameras. Table 1 summarizes currently active satellites. This table
will be updated as necessary. (NOAA-5 is planned for launch in July 1976.)

Orbits and products available

ATS-1 was launched on 7 December 1976 and positioned on the equator near
149° W at an altitude of 35,805 km (Pacific Ocean), with an orbital inclination
of less than 1^O. Perturbing forces have since that time produced an increase in
the inclination; consequently, ATS-1 now oscillates north and south of the equator
by more than 7° in a 24 hour period.

TABLE 1

CURRENTLY ACTIVE DIRECT READOUT SATELLITES

(as of June 1976)

Satellite	Launch Date	Orbit	Service*	Sensor*	Frequency (MHz)	Incl. (Deg.)	Period	Height Apogee/Perogee
NOAA-1	6 Nov. 1973	Polar	APT	SR	137.50/.62	102.0	116.2 min.	1509/1500 km
			HRPT	VHRR	1697.5			937/931 nm
			DSB	VTPR**	136.77/137.14			
NOAA-4	15 Nov. 1974	Polar	APT	SR	137.50/.62	101.7	115.0 min.	1457/1443 km
			HRPT	VHRR	1697.5			786/779 nm
			DSB	VTPR	137.14/136.77			
ATS-1	7 Dec. 1966	Geostationary	WEFAX	--	135.60	6.98	24 hrs.	35847/35763 km
								19356/19310 nm
ATS-3	5 Nov. 1967	Geostationary	WEFAX	--	135.60	5.36	24 hrs.	35885/35678 km
								19376/19265 nm
SMS-1	17 May 1974	Geostationary	WEFAX	--	1691.0	1.79	24 hrs.	358?1/35825 km
			--	VISSR	1687.1			19347/19345 nm
SMS-2	6 Feb. 1975	Geostationary	WEFAX	--	1691.0	0.60	24 hrs.	35804/35768 km
			--	VISSR	1687.1			19330/19311 nm

*Service/Sensor Acronyms

APT -- Automatic Picture Transmission
HRPT -- High Resolution Picture Transmission
DSB -- Direct Sounding Broadcasts
WEFAX -- Weather Facsimile
SR -- Scanning Radiometer

VHRR -- Very High Resolution Radiometer
VTPR -- Vertical Temperature Profile Radiometer
VISSR -- Visible/Infrared Spin-Scan Radiometer

**On Stand-by

ATS-3 was launched on 5 November 1967 and is presently positioned 70° W above the equator at an altitude of 35,800 km (Atlantic Ocean).

Antenna aim in Boulder, Colorado (105° W, 40° N) is:

for ATS-1   237° Azimuth
             25° Elevation

for ATS-3   132° Azimuth
             32° Elevation

These two satellites carry VHF transponders which are used to relay weather pictures obtained by the NOAA satellites. The NOAA pictures are computer processed on the ground with grid lines added and relayed to any ground station in view of the ATS satellites.

The SMS satellites will not be used in this experiment since the S-band receiver required is not yet operational.

The NOAA satellites are in so-called sun-synchronous polar orbits at altitudes of approximately 1500 km and periods of about 115 minutes. By inclining the orbits about 12° from a true polar orbit (see Fig. 1) the plane of the orbit precesses just 1 revolution per year, thus the plane of the orbit always points towards the sun and the satellite comes overhead in daylight about the same local time every day.
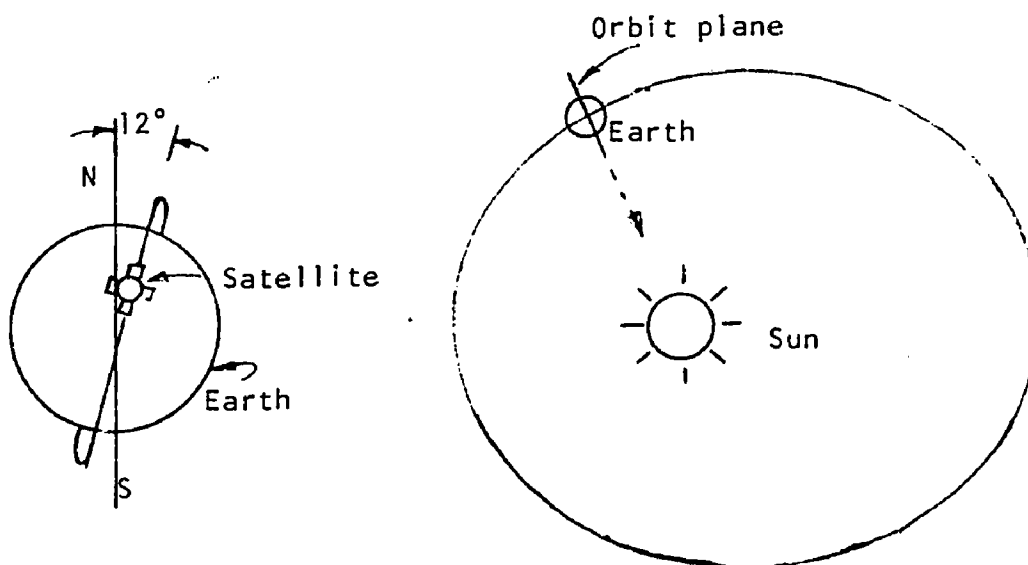

Fig. 1.   Sun-synchronous orbit.

Picture format and communication system description

Two different systems are in use: the old ESSA TV system whose format and scanning rate is used by ATS-1 and 3, and the newer NOAA system.

ESSA (ATS) system

One frame consists of 800 lines of video read out line-by-line at a horizontal scanning rate of 4 lines per second, thus it takes 200 seconds to transmit one frame. Fig. 2.
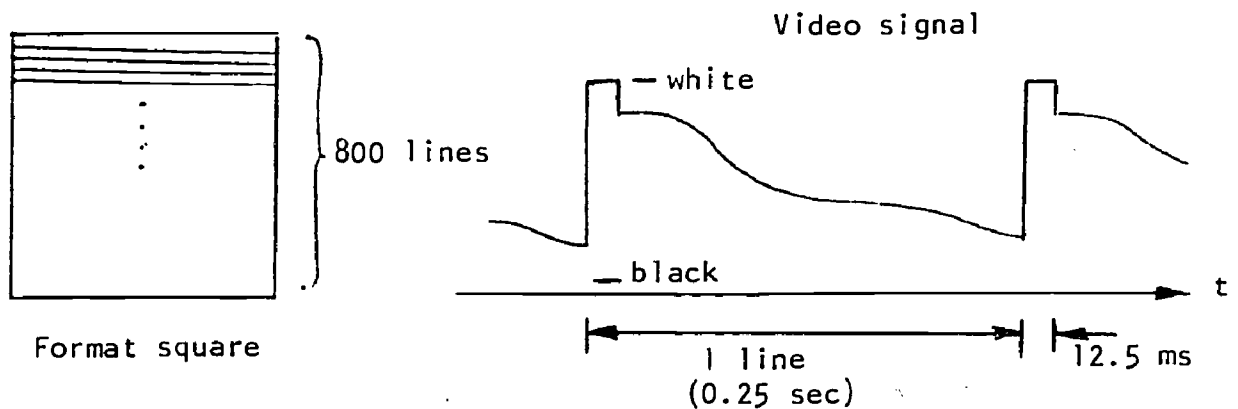

Fig. 2. Format and video signal ESSA system

The video frequency is from DC to 1600 Hz assuming a horizontal resolution of 800 dots. This video signal amplitude modulates a 2400 Hz subcarrier. The AM signal BW is from 800 to 4000 Hz (2400 ± 1600 Hz) and frequency modulates a VHF carrier in the 136 MHz channel with $\Delta f = 10$ kHz . The BW of the FM signal is approximately $2\Delta f = 20$ kHz . The 4 Hz horizontal scan is obtained from the 2400 Hz subcarrier (clock) by frequency division.
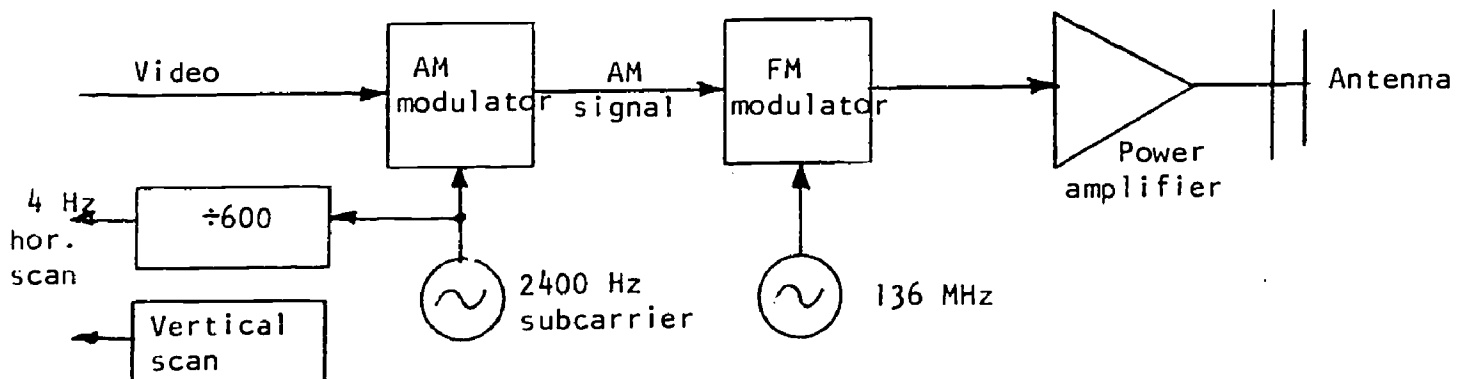

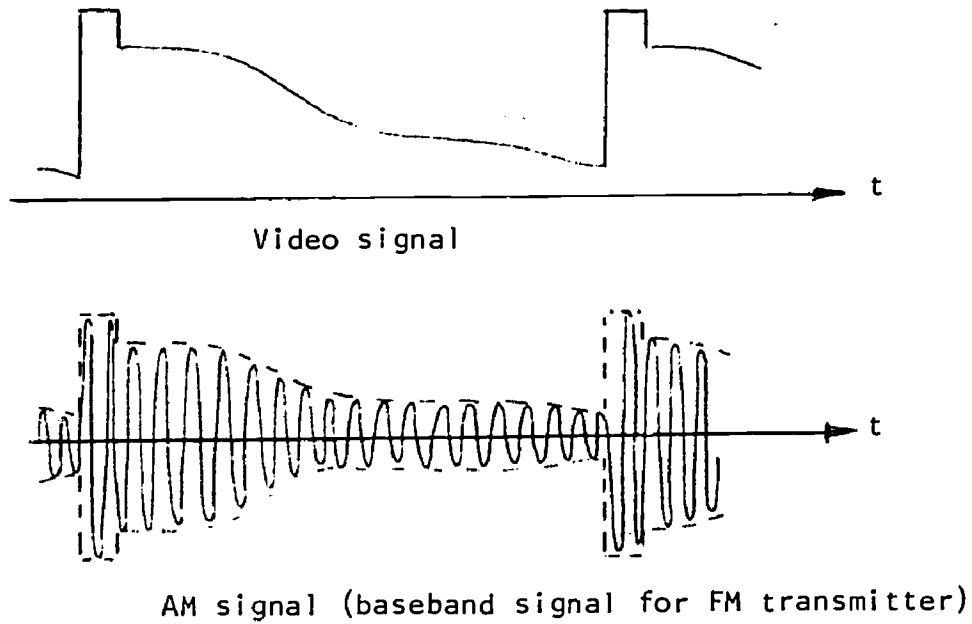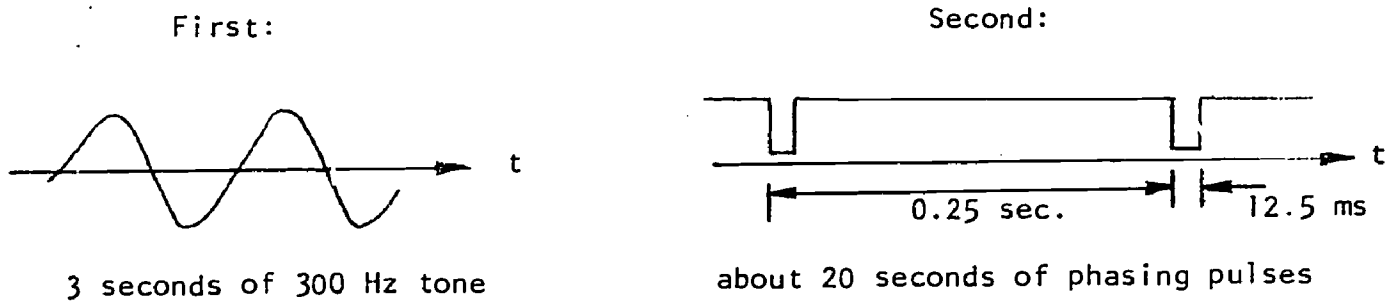Fig. 3. Simplified picture transmitter

Video signal

AM signal (baseband signal for FM transmitter)

Fig. 4. Typical video and AM signals

Just before transmitting the actual picture the following video is sent:

First: Second:

3 seconds of 300 Hz tone   about 20 seconds of phasing pulses

The 300 Hz tone indicates the end of one picture and the beginning of the next. The phasing pulses are used in the picture processor at the ground to align the start of the horizontal sweep with the edge of the picture.

NOAA system

The NOAA satellites use an optical system consisting of an infrared and a visible light sensor, a lens system, and a rotating mirror. The horizontal

scan is produced by the rotating mirror scanning the earth from horizon to horizon perpendicular to the orbital motion. Fig. 5.

scanning mirror

NOAA satellite at 1500 km altitude
Orbit perpendicular to paper

s ≃ 8000 km
(5000 mi.)
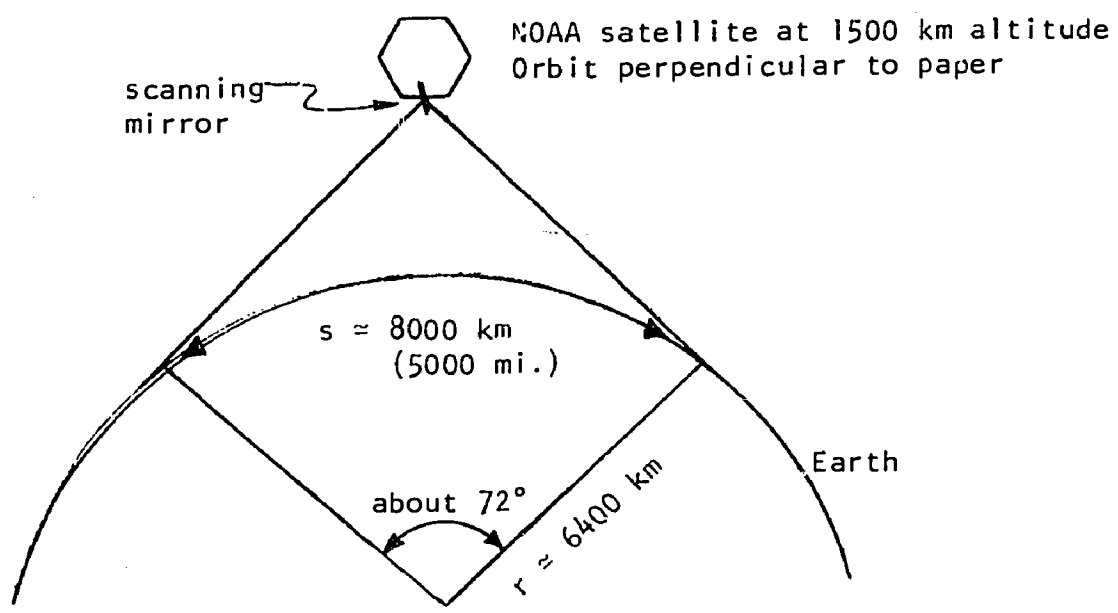
Earth

about 72°

r ≃ 6400 km

Fig. 5. NOAA scan system

The "vertical" scan is provided by the motion of the satellite in its orbit. The infrared data (IR) is transmitted to the ground during the scanning process (180° mirror rotation) while the visible channel (VIS) is tape recorded and played back and transmitted during the remaining 180° mirror rotation. Thus the infrared and visible lines are time division multiplexed (interleaved in time). In addition to the IR and VIS video picture data, there are synchronization pulses at the beginning of each line as well as calibration steps and so-called back and front porches. Fig. 6 shows the complete video signal for one IR and one VIS line. The scanning rate is one IR and one VIS line per 1.25 seconds, thus the scan is very slow. Also note that there is no beginning or end of the picture -- the satellite scans continuously. This composite video signal amplitude modulates a 2400 Hz subcarrier (just as in the ESSA system). The resulting AM signal frequency modulates a VHF main carrier.
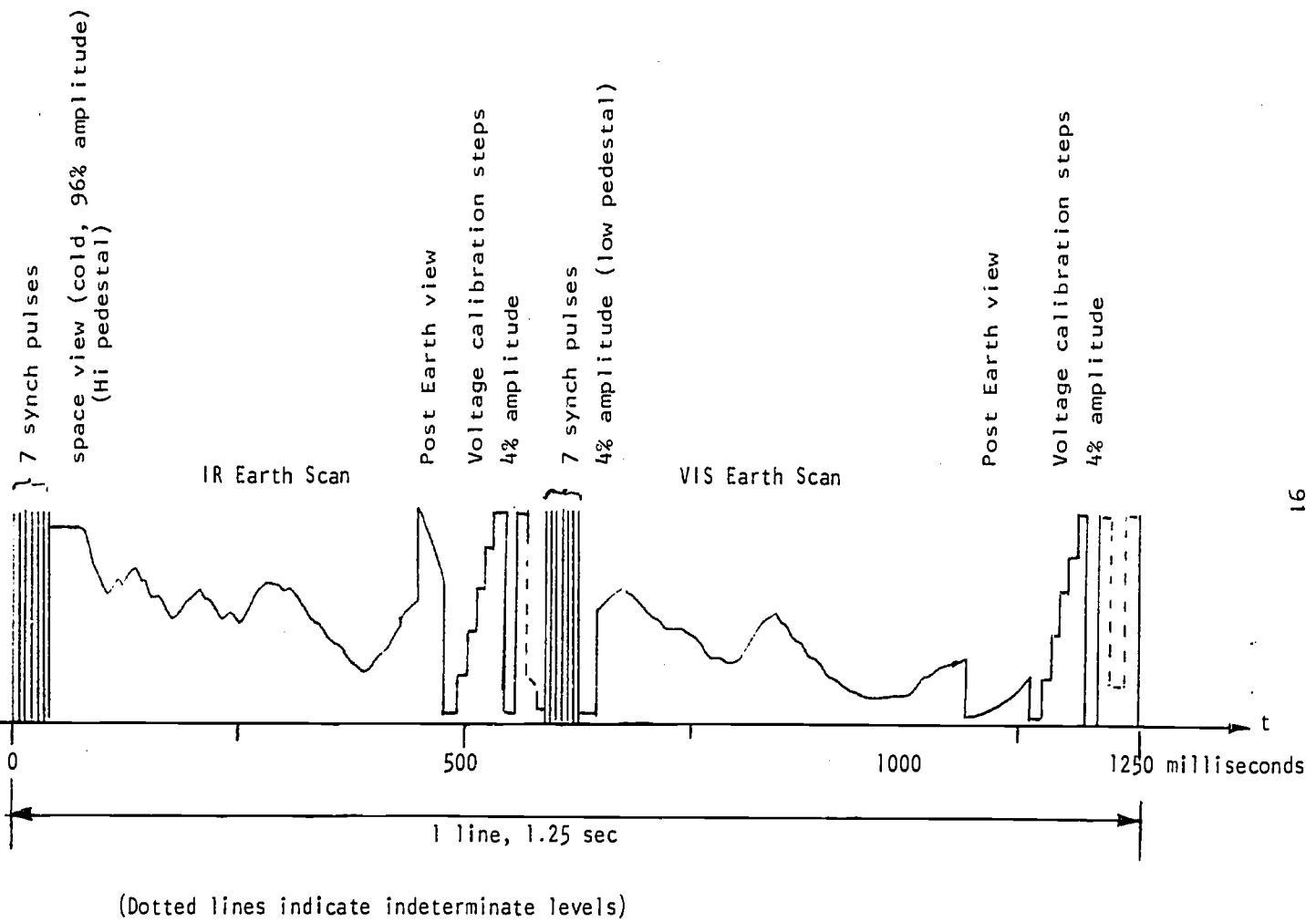
Fig. 6. Typical NOAA video signal

(Dotted lines indicate indeterminate levels)

1 line, 1.25 sec

Ground station

The ground station consists of a steerable helical antenna, antenna pre-amplifier, VHF-FM receiver, and a conventional audio tape recorder. Fig. 7.
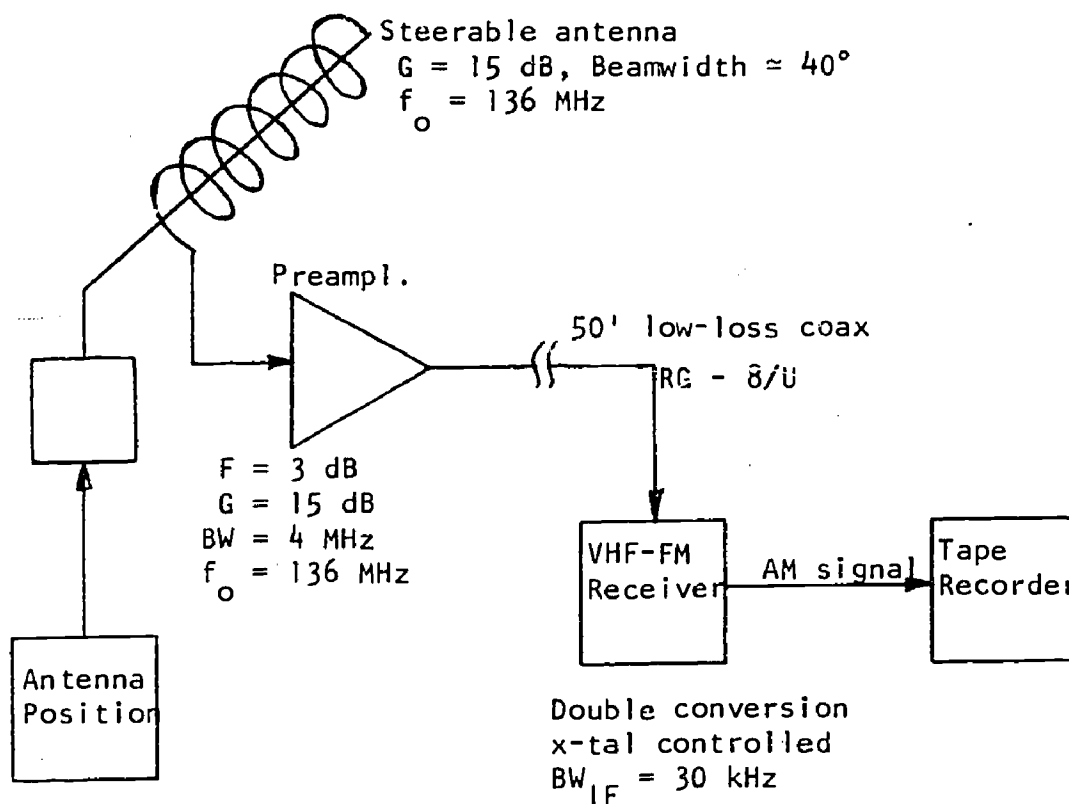

Fig. 7. Simplified block diagram of ground station

The audio output of the FM receiver is the AM subcarrier signal (frequency range 800-4000 Hz) and is tape recorded for later processing into pictures. The signal as transmitted by the satellite is linearly polarized; however, the polarization direction is unknown and will change with time (due to Faraday rotation in the ionosphere) by the time the signal reaches the ground station. Hence the need for a circularly polarized receiving antenna.

Picture processor

A simplified block diagram of the processor is shown in Fig. 8 for the ESSA system. Its main component is a specially built high resolution oscilloscope.
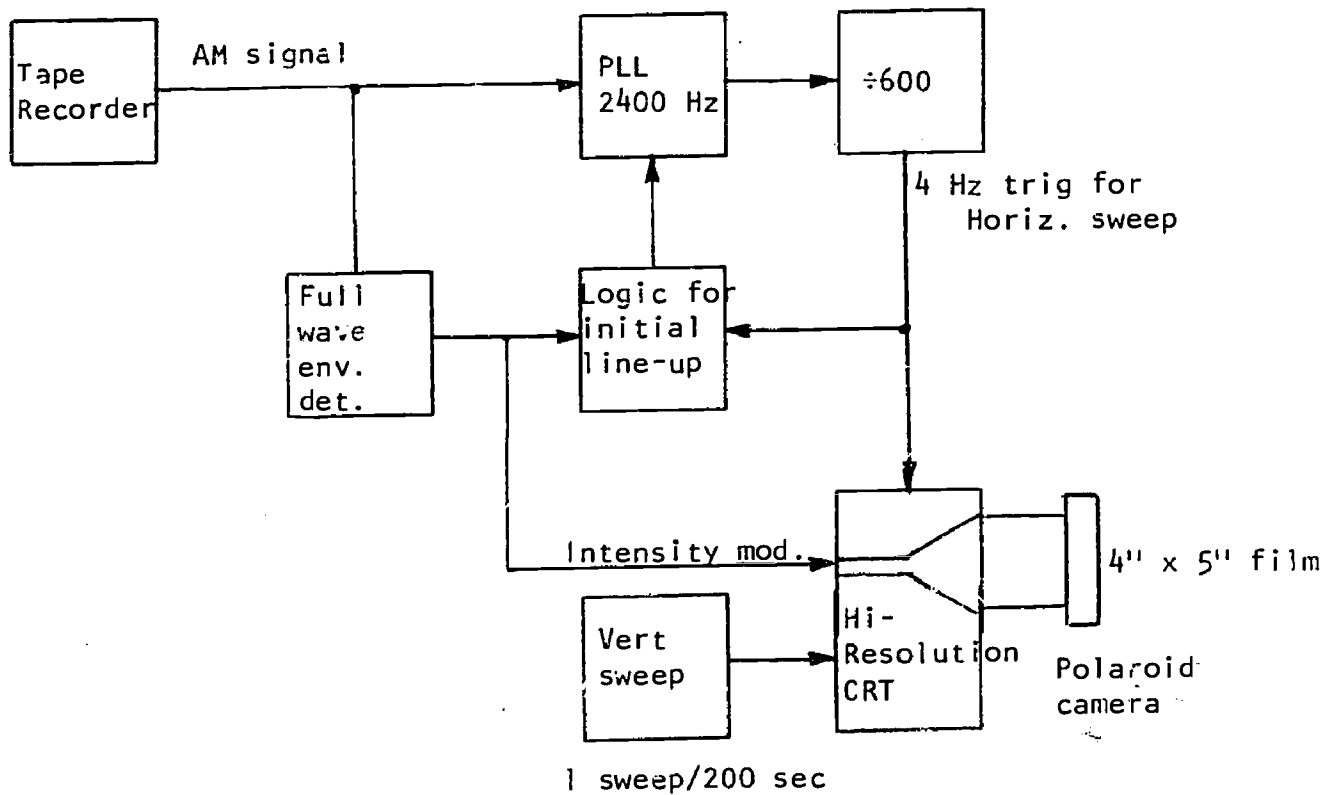
Fig. 8. Simplified block diagram of picture processor for ESSA system

Since the horizontal scan is derived from the 2400 Hz subcarrier in the satellite
a phase-locked-loop (PLL) is used to lock on to the subcarrier. By dividing
by 600 a 4 Hz horizontal trigger signal is obtained. This scheme eliminates
any effect due to errors in playback speed of the tape recorder. The AM signal
is also full wave envelope detected to recover the video signal which is used
for CRT intensity modulation. Initially, the 4 Hz trigger pulses may not occur
at the picture edge. By comparing the initial phasing pulses (20 seconds duration)
with the 4 Hz trigger pulses in the logic the PLL is automatically aligned so
that the edge of the picture occurs at the edge of the CRT. The picture is
recorded on 4" x 5" polaroid film.

Fig. 9 shows a simplified block diagram of the processor for the NOAA
pictures.

The logic detects the presence of the 7 synch pulses and either a low
(VIS) or high (IR) pedestal following the synch pulses, depending upon the
position of the IR/VIS front panel switch and generate horizontal synch pulses
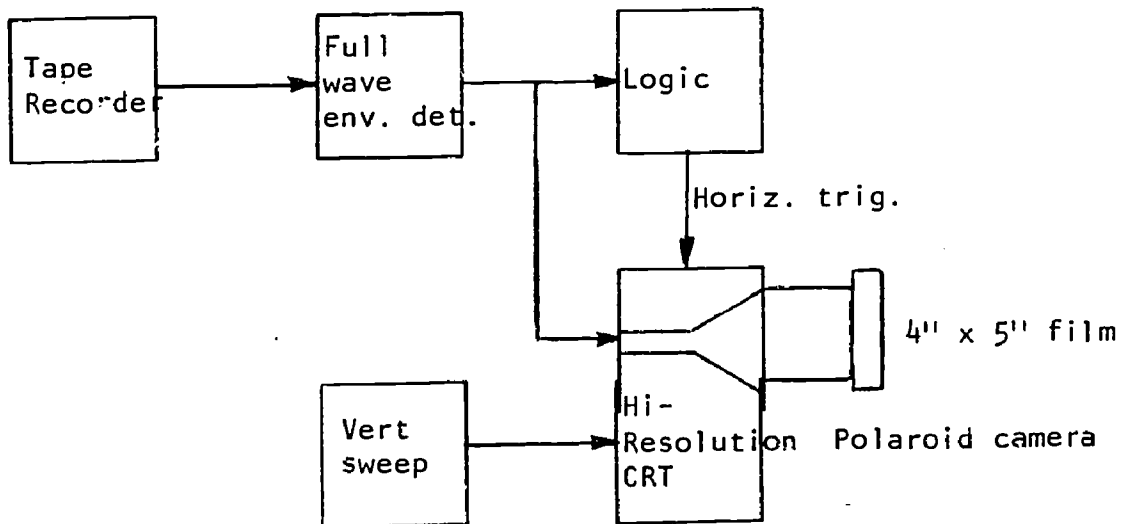
Fig. 9. Simplified block diagram of picture processor
for NOAA system

for the scope. The scope sweep speed is adjusted so that the beam sweeps across
the screen during one IR (or VIS) line for each synch pulse. Thus the video
signal will be demultiplexed and by playing the tape twice two pictures, one IR
and one VIS, will be obtained.

Operational details for both processors will be given as needed.

In this experiment it is most convenient to use the ATS satellites as no
tracking is necessary -- just point the antenna and turn on the receiver at the
current transmitting times. If you like to spend more time you may receive
the NOAA satellites. In this case, it is necessary to compute their position
versus GMT from orbital information. A computer program is available for this
together with current orbital parameters.

Procedure

1. Obtain current ATS or NOAA data, and set up the receiver and tape recorder.

2. Check the calibration of the receiver signal strength and frequency offset
    (discriminator) meters.

3. Receive a sequence of picture and record all pertinent data. Compare received carrier power with calculated value.

4. Process the picture information on the tape to produce 4" x 5" polaroid pictures.

APPENDIX

Down link (or power budget) calculation for ATS-3

Satellite

Signal power (carrier) = 5 watts = 7 dBW

Antenna gain relative to isotropic = 9 dB

Effective isotropic radiated power (EIRP) = 26 dBW (i.e., 400 watts)

Space

Free space "loss"* α = 32 + 20 log f + 20 log d = 177 dB

(f in MHz, d in km = 37,600 km)

Polarization loss   <u>3 dB</u>

Total loss   <u>180 dB</u>

Ground station

Receiver antenna gain relative to isotropic = 15 dB, thus

C = carrier power received = 26 + 15 - 180 = -139 dBW = -109 dBm,

and using an FM improvement factor of 10 dB, the received signal power

S = -139 + 10 = -129 dBW = -99 dBm.

Noise

Receiver noise figure = 3 dB   i.e., T_e = 300° K ,

Sky temperature at 136 MHz T_s = 600° K , thus

Noise power/Hz = $k(T_e + T_s)$ = k(900° K) $\simeq 10^{-20}$ watts/Hz

IF BW = 30 kHz, Noise power = N = $k(T_e + T_s)BW$ = 3×10^{-16} watts

                                 = -155 dBW

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*ITT Ref. Data for Radio Eng., 1975

Finally:

$$\frac{C}{N} = -139 - (-155) = 16 \text{ dB}$$

and $\quad \frac{S}{N} = -129 - (-155) = \underline{\underline{26 \text{ dB}}}$