Instructor`s Manual

to accompany

Communication Systems Modelling

with

tims

Volumes A1, A2, D1, & D2

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INTRODUCTION

this manual

This ***TIMS Instructor`s Manual*** is intended to help you, the Instructor, use and supervise each experiment. It contains notes on each experiment, as well as answers to most Tutorial Questions.

Each of the experiments described in the four volumes of ***Communications System Modelling with TIMS*** [1] .

Sufficient details are supplied, including background information, for the block diagram to be modelled and adjusted to perform as required.

Some experiments are written with the expectation that earlier experiments will have already been completed. This is stated under the *prerequisites*.

Most experiments include some suggested measurements. Depending upon the level of your course work you may find it necessary to add your own list of supplementary measurements.

the TIMS laboratory

A well equipped laboratory will have all TIMS SYSTEM UNITS interconnected via the TIMS TRUNKING SYSTEM, with one TIMS SYSTEM UNIT serving as the MASTER, the remainder being SLAVES.

Some experiments require that the MASTER generates signals which are sent to the SLAVES via TRUNKS.

For building very large systems a TIMS Junior can be added. This will expand the capacity from 12 modules (of the standard TIMS 301) to a total of 20 modules.

100 kHz MASTER clock

Unless there is a special reason for not doing so, all TIMS SYSTEM UNITS in the laboratory should be synchronized to the 100 kHz clock in the MASTER TIMS SYSTEM UNIT.

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1 sometimes referred to as the 'Student Text'.

messages

Many experiments require signals to be derived from 'messages'.

There are many sources of messages, including:

analog messages

1. a SPEECH MODULE will be introduced during 1999
2. AUDIO OSCILLATOR
3. analog output from the VCO.
4. the 2.083 kHz message from the MASTER SIGNALS module.
5. two-tone signals (add any two of the above)
6. band limited recorded speech (see below)
7. the lowpass filtered analog output from a SEQUENCE GENERATOR.
8. DC (to suggest a telemetry signal in TDM say).

digital messages

Almost exclusively digital messages will come from a SEQUENCE GENERATOR module. However, there are exceptions.

bandlimited speech

We recommend that you provide yourself with a source of taped speech, as this is always useful for qualitative observations, and is essential for some experiments (two or more tracks are required for TDM and FDM).

Speech can be obtained by recording radio programs, although such signals tend to have wide variations in level. A better solution is to make a voice recording under studio conditions, from suitable non-copyright text. A segment of only a few minutes is adequate. This should then be copied onto a second cassette, and repeated for as many times as required, so that the second cassette will last for the duration of the experiment. By repeating the text the student soon becomes bored with the contents, and can then concentrate on the experiment.

Two or more speech channels are useful for multiplexing experiments. The topics chosen for each track should differ as widely as possible. This enables the student to identify, from the context, which channel is being recovered. If you are preparing your own tapes then readings from technical, fictional, sporting texts and so on can satisfy this requirement.

The speech *must* be bandlimited to the range 300 to 3000 Hz. Attenuation of frequencies above 3 kHz is essential for many experiments (eg TDM and FDM). A TIMS TUNEABLE LPF can be used for this purpose whilst making the recording.

To cut the low frequencies a -3 dB corner at around 300 Hz is suggested. This can be achieved with an external RC network inserted in the signal path during the recording. Unless the low frequencies are attenuated *as well as* the high frequencies (above 3 kHz) the intelligibility will suffer.

tutorial questions

The initial aim of the Tutorial Questions was to ask questions which could be answered experimentally, during the experiment. Thus purely analytical questions were to be avoided.

As experiments were added this philosophy was sometimes forgotten, and analytical questions crept in. These questions have not been edited out, but answers are not always supplied.

note taking

No advice has been given about note taking. This does not mean that note taking is unimportant. You should have a definite policy as to how you want the results of each experiment recorded. Students should be made aware of this.

laboratory programs

TIMS is capable of supporting courses in communications at all levels. These courses are typically given in two, often independent, strands; namely analog and digital. However, it is perfectly feasible to give a general course on communications without introducing this split. What ever the course structure, TIMS can support the theoretical side of any such course with related experiments.

Many instructors will prefer to write their own laboratory material, extracting those parts of this Text which suit their purposes. Others might prefer to build the theoretical side of their course around a selection of the experiments contained herein.

CONTENTS

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INTRODUCTORY ANALOG EXPERIMENTS

Introduction to modelling with TIMS

It is recommended that this Chapter be read by all students prior to attempting their first experiment, be it analog or digital.

Modelling an equation

TRUNKS

Depending on the quality of your students, and the time available, you may feel the experiment is long enough already. Otherwise, it is instructive to supply a bandlimited speech signal at TRUNKS. Preparation of such a signal is discussed in the ***Introduction*** to this ***Instructor`s Manual***. Alternatively you can use a SPEECH [1] module.

There are then two options suggested; you can think of others.

1) to the speech add a single tone from an AUDIO OSCILLATOR. The tone frequency should lie within the speech band - say 1 kHz.

You can set the relative level of the speech and tone to any value which you find convenient; try the experiment, and see which you prefer.

When this corrupted speech signal is used as the input to the nulling model it should be possible to null the tone from the output. This adjustment will put a 'hole' in the speech near the tone frequency, but the speech will still be intelligible.

2) add a second tone to the speech (or replace the in-band tone), this one being *above* the filter slotband. This would be removed when the LPF of the HEADPHONE AMPLIFIER is switched in, giving a demonstration of noise removal when the wanted and unwanted signals are non-overlapping in frequency.

oscilloscope triggering

As stressed both in the introductory chapter of the student text, and in this experiment, it is important for students to develop an early appreciation of the need to choose a 'suitable' signal for oscilloscope triggering.

A 'suitable' signal must at least be of constant amplitude. That is, it should not change as the various adjustments are made to signal levels throughout the model. This obviates the need for constant re-adjustment of the triggering circuitry.

[1] available 1999

Typically this signal will be periodic. The oscilloscope trigger controls must be set to accept this external trigger signal, and preferably switched to 'auto mode' if available.

answers to tutorial questions

Q1 the depth of null needs to be described with respect to a reference. This is generally the amplitude of the 'wanted' term. Here there is no wanted term. So why not use the *sum* of the two components at the ADDER output as the reference. This makes the calculation easy, since after measuring the null (the *difference*), just flip the 180 degree switch on the PHASE SHIFTER front panel, and turn the cancellation into an enhancement, which displays the *sum* as required.

It is usual to express this in decibels, thus:

$$\text{depth of null} = 20 \log_{10} (sum / difference)$$

If the two signals out of the ADDER are $V_1.\cos\omega t$ and $V_2.\cos(\omega t + \alpha)$, then when the signals are subtracted, the difference amplitude is:

$$difference = \sqrt{[(V_1 - V_2 \cos\alpha)^2 + (V_2 \sin\alpha)^2]}$$

$$sum = \sqrt{[(V_1 + V_2 \cos\alpha)^2 + (V_2 \sin\alpha)^2]}$$

Setting $V_1 = 1.01\ V_2$ and $\alpha = 1$ degree, then depth of null is close to 40 dB

Q2 the PHASE SHIFTER introduces different phase shifts at different frequencies. To make an analog wideband constant phase shift network is hard enough, but to make it adjustable as well is *very* difficult (impossible ?). Periodic signals in the noise, other than the one for which the cancellation has been achieved, are unlikely to arrive at the ADDER exactly out of phase, so would not cancel (although may be reduced in amplitude). Circuit noise, being wideband, will likewise not cancel over a wide bandwidth.

Q3 the method, without refinements, would be approximate only. Measure the total power out when there is no cancellation (one ADDER input removed), then cancel the fundamental, and measure remaining power. Assumes random noise power is even lower than the harmonics, otherwise is measuring signal to signal-plus-noise-and-distortion ratio.

Q4 since the phase shift introduced by the PHASE SHIFTER varies with frequency, the null will not be maintained. Refer to an earlier question.

Q5 in principle, yes. But the PHASE SHIFTER must compensate for small phase shifts within each module, and so itself may need to introduce slightly more or less than exactly 180^{o}. This the INVERTING AMPLIFIER would not do.

DSBSC generation

TRUNKS

The experiment does not require any TRUNKS signals. However, if you like, you could send speech, so that students can see what a DSBSC signal looks like when derived from such a message. Preferably they could have SPEECH modules.

Task T5 requires the creation of a text-book-like display of a DSBSC. This means that *both* the envelope *and* the DSBSC itself are stationary on the screen. Use a message which is a sub-multiple of the 100 kHz carrier. This can be done with the 2.083 kHz sinusoidal message from the MASTER SIGNALS module (this is a 48:1 ratio).

There is also an 8.333 kHz (ratio 12:1) TTL signal at the MASTER SIGNALS module. The TTL can be used instead, but its fundamental must first be extracted with the TUNEABLE LPF. Overload of the analog filter with a TTL input is not a problem - as long as the output is near-sinusoidal it is acceptable. Switch the MULTIPLIER to AC coupling, else the DC from the TTL will result in a carrier term at the MULTIPLIER output (and the output signal will no longer be a DSBSC).

answers to tutorial questions

Q1 The product:

$$V_1.\cos\omega_1 t \text{ and } V_2.\cos\omega_2 t \ = \ k.V_1.V_2.½[\ \cos(\omega_1 - \omega_2)t + \cos(\omega_1 + \omega_2)t\ ]$$

where ‘k’ is a property of the MULTIPLIER.

In the multiplier module ‘k’ has been set so that, when V_1 and V_2 are at the TIMS ANALOG REFERENCE LEVEL, so also will be the peak amplitude of this product.

Thus, working in peak values, we want

$$k.V_1.V_2.½\ [1 + 1] \ = \ V_1$$

which gives:

$$k \ = \ ½$$

Q2 since a DSBSC contains more than one component the question is meaningless - unless, perhaps, if the DSBSC was periodic. See next two questions. Of interest is often the frequency of its suppressed ‘carrier’.

Q3 the FREQUENCY COUNTER counts level crossings in one direction only in a given time interval (set by front panel switch). This level is above zero, but well below the TIMS ANALOG REFERENCE LEVEL. The zero crossings of a DSBSC signal are not uniform. You should consider how you might attempt to prove this, or otherwise define the zero crossing locations. The DSBSC signal is not periodic - at least, not in a time interval comparable with the message period, let alone the carrier period. A quick answer is that you don`t know what the counter will display, but it surely won`t be the carrier frequency. Secondly, it will probably vary from count to count.

Q4 see previous answer.

Q5 three pairs of DSBSC components of relative amplitude A_1, A_2, and A_3. Absolute amplitudes unspecified.

Q6 analytically it is not necessary. But if oscilloscope displays are to be compared with ‘normal’ expectations, then it *is* necessary. See the experiment entitled *Envelopes* in Volume A1.

Amplitude modulation

If you have them, WIDEBAND TRUE RMS METER modules could be useful (eg, see Tutorial Question)

TRUNKS

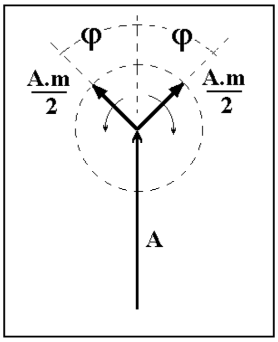
Send speech; but if students have SPEECH modules this is unnecessary.

answers to tutorial questions

Q1 while watching the envelope, and the trough-to-trough distance Q, increase ‘m’ from below unity to above unity. If you imagine ‘Q’ changing sign as m passes through unity, then it becomes negative, and the formula can still be used.

Q2 this is not easy to explain in a few words. Set it up with TIMS, and work out an explanation yourself !

Q3 the phasor diagram represents an AM signal $y(t) = A(1 + m.\cos\mu t)\cos\omega t$:



This can be represented in phasor form as in the figure.

When $\varphi = 180$ degrees the resultant gives the trough amplitude, which equals Q/2.

$Q/2 = A - Am/2 - Am/2$

When $\varphi = 0$ degrees the resultant gives the peak amplitude, which equals P/2.

$P/2 = A + Am/2 + Am/2$

Substitute for P and Q, from these two equations, into the right hand side of eqn.(9) and show it is equal to m.

Q4 the AC/DC switch is a dual switch. It controls signals *into* the MULTIPLIER. It passes AC only in the ‘AC’ position, but AC *and* DC in the ‘DC’ position.. In the present vase, removal of the DC term in the message eliminates the carrier term - see eqn(1) - so the output signal changes to DSBSC.

Q5 in the phasor diagram the three amplitudes will be ½, 1, and ½ volt peak. The total power in 1 ohm will be 0.75 watt. Thus the rms meter will read $\sqrt{(0.75)}$ volt.

Q6 a reversal of the phase of the carrier term.

Envelopes

TRUNKS

No signals specified.

answers to tutorial questions

Using phasors to predict envelope shapes is a useful skill to acquire. The technique is often not fully appreciated until a non-standard case is examined, such as that of Q2.

Q1, 2 draw the phasors at various positions throughout a period of the message. Measure their resultant. This gives a point on the envelope amplitude versus time curve.

If the carrier is drawn vertically, then, for the *special case* of carrier and resultant being co-linear (amplitude modulation), the graphical construction of the envelope is simplified by projecting a parallel line from the resultant peak across to the envelope curve.

But if the DSBSC resultant is *not* co-linear with the (vertical) carrier phasor, this simple construction technique (projecting across) is invalid. The resultant amplitude must be measured, and used for the envelope amplitude at that time, but the point on the envelope curve cannot be located by simply projecting across horizontally.

Q3 radio signals can arrive at an antenna via one or more paths. If more than one this is called multi-path reception. If the paths are of different lengths, the resultant will be the phasor sum. Particularly in the case of ionospheric paths these all change with time (measured in seconds). So the received signal strength varies with time.

It can be shown that components even very close in frequency (as are sidebands in a narrow-band signal) will suffer different fading effects - hence the term 'frequency selective' fading.

It is possible for the carrier of an AM signal to fall (even if only momentarily) to zero amplitude, while sideband amplitudes remain finite. Reference to the relevant phasor diagrams will show that the envelope suffers *severe* distortion.

See 'multi-path' propagation.

Envelope recovery

TRUNKS

No TRUNKS signals are specified. If no SPEECH modules then supply it if you consider it useful. You could also send signals with envelopes to be identified. These could be derived from speech, or tone (or tones), whose frequencies must be determined.

answers to tutorial questions

Q1 & 2 for a fullwave rectifier in the ideal envelope recovery circuit the requirement is that:

$$\mu < (2\omega - \mu)$$

For a half wave rectifier it is that:

$$\mu < (\omega - \mu)$$

This is derived by reference to Figure 2A of the experiment.

In discussing envelopes, and distortion in the output of envelope recovery circuits, one should be clear as to what is required.

By definition there is no distortion of the *envelope* recovered with an ideal rectifier-filter combination (assuming an appropriate filter).

If one is actually looking for the *message*, then that is dependent on the nature of the signal. Thus, for an AM signal, it is necessary that $m < 1$. If $m > 1$, then the envelope will be a distorted version of the message. Thus, in this case, the output from an envelope detector will be a distorted version of the *message*, but not a distorted version of the *envelope*.

The square law detector can be analysed by considering its transfer function to be:

$$e_{out} = a_1 e_{in} + a_2 e_{in}^2$$

For the case of the input being the AM signal

$$y(t) = (1 + m.\cos\mu t)\cos\omega t$$

then the audio output terms (from an appropriate LPF, and dropping the scaling factor $\frac{1}{2}a_2$), are:

$$LPF\ output = 1 + \frac{m^2}{2} + 2m\cos\mu t + \frac{m^2}{2}\cos 2\mu t$$

Ignoring the DC term, then the amplitude ratio of the wanted (fundamental) to the only unwanted (2nd harmonic) term is:

$$ratio = \frac{4}{m}$$

This, for m = 1, is not very good, but for smaller values of 'm' it is acceptable.

Q3 apart from a scaling factor, the unwanted components lie near 2ω for the full wave rectifier, but near ω for the half wave. Thus the filtering requirements are different. This is perhaps of little significance when ($\omega >> \mu$); but otherwise may be important.

Q4 radio signals can arrive at an antenna via one or more paths. If more than one this is called multi-path reception. If the paths are of different lengths, the resultant will be the phasor sum. Particularly in the case of ionospheric paths these all change with time (measured in seconds). So the received signal strength varies with time.

It can be shown that components even very close in frequency (as are sidebands in a narrow-band signal) will suffer different fading effects - hence the term 'frequency selective' fading.

It is possible for the carrier of an AM signal to fall (even if only momentarily) to zero amplitude, while sideband amplitudes remain finite. Thus the envelope suffers severe distortion.

Thus the output from an envelope recovery circuit delivers a non-linearly distorted message (intelligibility is severely impaired), whereas the output from a synchronous demodulator exhibits only linear distortion (effect upon intelligibility may be mild).

SSB generation - the phasing method

TRUNKS

Speech is interesting. Send it if students do not have SPEECH modules.

answers to tutorial questions

Q-unnumbered. There is no reliable method of positively identifying, in the time domain, a signal as either SSB or DSBSC when derived from speech.

Q1 put the PHASE SHIFTER in the path to the other MULTIPLIER. There are others. Try not to implement a change which makes two reversals, thus cancelling each other.

Q2

Q3 frequency division multiplex (FDM) systems were used in vast quantities for analog telephone systems before being superseded by time division multiplex (TDM). One such system used channels spaced by 4 kHz in the range 60 to 108 Hz. Channel separation was accomplished with bandpass filters, the cost of which was low due to the quantities involved. There were thus filters located either side of the TIMS 100 kHz carrier, and these were used by TIMS as LSB and USB filters. These filters are no longer easily obtainable.

Q4 by eye one would have to detect when a circle had degenerated into an ellipse. Say a difference of 5% in the major and minor axes ? Not good enough to detect 1 degree errors, but OK for a quick check for major problems.

Q5 no ! The Hilbert transformer is a mathematical transform which conceptually is a single-input, single-output device. It has an infinite bandwidth, extending down to DC. Practical realizations have a finite bandwidth, which does *not* extend down to DC.

The QPS is composed of two complementary devices, each with a single-input and single-output. The phase shift through each device varies (differently) with frequency, but at any one frequency these phase shifts sum to 90^0 (± a small, typically equiripple, error).

Q6 the arrangement of Figure 4 would be a quick check for serious errors. The most realistic check is to use it in an SSB generator (or receiver) and to measure sideband suppression (or rejection).

Q7 the requirement here is to find the ratio of the sum and difference of the two phasors V_1 and V_2, where they are α degrees off cancellation.

$$\text{sum magnitude} = \sqrt{[V_1 + V_2 + 2.V_1.V_2.\cos(\alpha)]}$$

$$\text{difference} = \sqrt{[V_1 + V_2 - 2.V_1.V_2.\cos(\alpha)]}$$

$$\text{suppression ratio} = 10 \log_{10}(\text{sum}^2 / \text{difference}^2)$$

For the case of a phase error only, $V_1 = V_2$ and this simplifies to:

$$SSR = 10 \log_{10} [\frac{2 + 2\cos\alpha}{2 - 2\cos\alpha}]$$

$$SSR = 10 \log_{10} [\frac{(1 + \cos\alpha)^2 + \sin^2\alpha)}{(1 - \cos\alpha)^2 + \sin^2\alpha)}] \; dB$$

which further simplifies to

$$SSR = 20 \log_{10} [\cot(\frac{\alpha}{2})] \; dB$$

For α = 1 degree this gives SSR = 41.2 dB

Q8 let the signal be y(t), where:

$$y(t) = 2\alpha\cos(\omega + \mu)t + \cos\omega t$$

where the amplitude ratios are 2α/1 This can be expanded into:

$$y(t) = (1 + \alpha.\cos\mu t).\cos\omega t - a.\sin\mu t.\sin\omega t$$

$$= a(t).\cos(\omega t + \beta)$$

where:

$$a(t) = \sqrt{(1 + 2.\alpha.\cos\mu t)}$$

Using the binomial theorem, or otherwise, this, for small α, approximates to:

$$a(t) = (1 + \alpha^2/4) + \alpha.\cos\mu t + \alpha^2/4.\cos2\mu t$$

Amplitude ratio of unwanted to wanted AC components is = α/4.

What size would 'α' need to be for the presence of the second harmonic to be just noticed ? Say 0.1 ? Set it up on TIMS and find out !

Q9 the spectrum is two lines, same amplitude, similar frequencies. In the time domain this looks exactly like a DSBSC signal. The envelope peaks will be separated by a time t_0, where:

$$t_0 = 2.\pi / \delta f$$

and δf is the frequency difference between the two audio tones.

Q10 it is useful for generating circles on an oscilloscope screen; perhaps for a circular time base ?

Product demodulation - sync & async

TRUNKS

You should generate the signals to be demodulated and send them via TRUNKS. Remember the students have been told each signal is based on a 100 kHz carrier, of which each TIMS 301 has a copy. This will become the ‘stolen carrier’ the students require for demodulation.

Required signals are:

DSBSC: choose any suitable message, typically speech, or a single tone. A two-tone audio is legitimate, if students are sufficiently experienced to cope with it. It will 'look' unusual, but its behaviour with respect to phase adjustment must reveal it as true DSBSC.

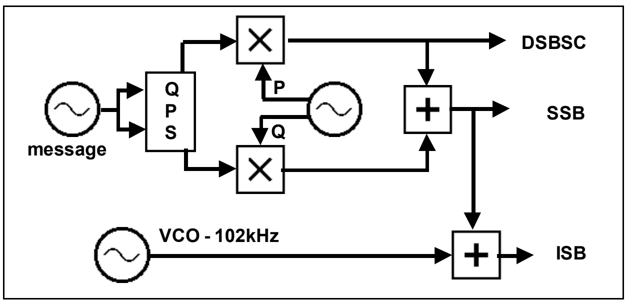
SSB: use a phasing type generator with speech as the message. If you prefer an USSB with a 1 kHz tone message, then the signal can be simulated with a 101 kHz output from a VCO.

A 'trick' message is a two-tone audio, which makes the SSB *look* like DSBSC, yet behaves like SSB (eg, recovered message amplitude is insensitive to local carrier phase). Do this only if the students are sufficiently experienced to cope with it. It can be simulated with two VCOs, say on 101 and 102 kHz.

ISB: you can make this with TWO generators, each of the phasing type, and using different speech messages.

Alternatively, ONE genuine SSB generator for the upper sideband (say), derived from speech, and for the other sideband ADD a sinewave, say 99 kHz, from a VCO. This simulates a lower SSB derived from a 1 kHz tone.

The ISB signal will show up the shortcomings of the synchronous demodulator under study, since it cannot differentiate between upper and lower sidebands. The true SSB demodulator will be examined later, in the experiment involved with SSB demodulation (and CSSB).



three possible TRUNK signals

You can think up other signals if you want to add your own personal touches. For example, a good ‘trick signal’ is:

CSSB - compatible SSB. This has not been defined in any experiment to date, but has probably been met in formal lectures ? This is made by adding a large carrier component to an SSB. If the SSB was derived from a single tone, then CSSB looks like AM with a small depth of modulation. Its message can be recovered with an envelope detector (for small SSB/carrier amplitude ratio the distortion won`t be noticed) and a synchronous demodulator, but the output from the latter is not dependent upon the local carrier phase. This is a 'trick signal' and should test the student`s understanding.

note that the CSSB signal described above does not need an SSB generator for its production - just ADD a small sinusoidal component from a VCO (on say 102 kHz), to a large 100 kHz carrier. Adjust relative amplitudes to make a reasonably undistorted sinusoidal envelope (say in the amplitude ratio 1:3 or less).

The student should be able to:

- identify each of the signals
- identify the messages each carries. For the case of the two tone message it is possible to identify the frequency of each tone (for example, filter off the lower of the two with the TUNEABLE FILTER; then recover the envelope of the two tone message to obtain the difference frequency).
- write an expression for each of the received signals

no TRUNKS ?

It is important to have a TRUNKS system for this experiment. To expect the student to generate the unknowns as well as identify them is not satisfactory. The unknown element is lost; and setting up would take an unnecessarily long time, and if badly done could lead to undesirable results.

Measuring δf

Easy in principle - multiply the 100 kHz reference with the VCO output. Filter with the RC filter in the UTILITIES module. Measure the frequency of the filter output. But ......

problem #1: the analog input of the COUNTER is AC coupled ! So put the filter output through the COMPARATOR and connect to the counter TTL input. This works fine.

problem #2: the counter needs a full second to respond for frequencies 1 Hz and above; even longer for smaller frequencies. One soon finds the method unsatisfactory as δf becomes small.

solution: the two sliding traces on the oscilloscope is by far the preferred method of observing and estimating the frequency difference.

answers to tutorial questions

Q1 try it ! Different observers tend to report different opinions, and text books have quite differing views.

error	*SSB*	*DSBSC*
0.1 Hz	not observable	no observable distortion, but amplitude will momentarily fall to zero every 5 seconds
10 Hz	probably not observable	horrible?
100 Hz	still intelligible especially if speech is translated high	horrible!

Q2 the synchronous demodulator as examined is not a true SSB demodulator, although it can demodulate SSB if an SSB is present alone.

The synchronous demodulator examined in the experiment has a window either side of the local carrier frequency, of width equal to *twice* the bandwidth *B* of the associated LPF; that is, from $(\omega - B)$ to $(\omega + B)$.

A true SSB demodulator must be able to 'look' at one side of the carrier, whilst ignoring any signal (including noise) on the other. It must be able to *select* sidebands.

Q3 a small DC would appear at the demodulator output. This would not be detected by ear !

Q4 refer to the diagram. The top receiver uses a synchronous demodulator, whereas the lower one uses an envelope detector. Assume the input signal is free of any significant noise. Provided there is only one input signal present then there would be no significant difference between the two audio outputs as observed by a *listener*. Instrumentation might detect a difference in signal-to-noise ratio.

There is a *significant* difference between the two receivers, however, when it comes to selectivity.

Suppose the bandwidth of the LPF was 0-3 kHz, and that the original AM signal had 3 kHz wide sidebands.

Consider the presence of a second signal of similar bandwidth, but 10 kHz higher in frequency.

- the *upper receiver* would frequency translate it both *up* and *down*. The *down* products would not pass through the LPF, since they would be in the range 7 to 13 kHz. The audio output from the wanted channel would remain unchanged. The bandwidth of the receiver - assuming linearity - is ±B, the audio filter bandwidth.
- the *lower receiver* would be operating on the *linear sum* of the two signals. The envelope of this combination would not be the sum of the individual envelopes. It would bear little (no ?) resemblance to either envelope (as an analysis will show). Thus the audio output would be unintelligible. The bandwidth of the receiver is infinite !

The superiority of the synchronous demodulator under these conditions is demonstrated. But this can be down-graded if there is any non-linearity in its circuitry (typically in the multiplier), since this will cause intermodulation products to be generated, some of which could pass through the LPF.

There is no inherent selectivity in the lower receiver. The audio filter is of no help in separating signals at RF. This would have to be done with a pre-detector filter, which would of necessity be an RF bandpass type.

Q5 the receiver should ideally be linear. (a) if operating at the TIMS ANALOG REFERENCE LEVEL the addition of a small amplitude signal at 90 kHz should not be noticed. (b) & (c) eventually, as its amplitude is increased, overload of some circuit will occur (probably the multiplier), new components will be generated, some of which will be observed.

Q6 see answer to similar question in the experiment entitled *Envelope recovery*. The synchronous demodulator would suffer from *linear* distortion - individual frequency components would be of the wrong amplitude (audible), and wrong phases (not detectable by ear, although the waveform *shape* would be incorrect). Importantly there are *no new frequency components* (which would occur with non-linear distortion). With speech the distortion could be of minor significance

Q7 see the chapter entitled *Introduction to modelling with TIMS*, under the heading *multipliers and modulators* (page 13).

SSB demodulation - the phasing method

TRUNKS

SSB, ISB, and DSBSC signals are required.

1. the SSB should be derived from speech.
2. the ISB should have at least one sideband derived from speech
3. the DSBSC should be derived from speech, with a sinewave (from VCO) added to one sideband to simulate an interfering carrier. Its amplitude should be of similar magnitude to the DSBSC itself, or otherwise as you decide.

It does not matter if the speech is the same for all, although for proof-of-identification purposes separate messages would be preferred. But this will require more modules at your end.

An economical method of producing these signals is to use an arrangement based on that used in the previous *Product Demodulation* experiment. It needs *three* ADDERS. Thus:

1. USSB from the above generator
2. one side of the ISB is the above SSB, into an ADDER, together with the other supplied by a VCO in the range say 97 - 99 kHz.
3. DSBSC can come from within the phasing generator, into an ADDER, into which is connected the interference from a VCO (the same as for ISB) on top of either sideband.

If you have more modules available then it would be preferred that the three signals looked a little less alike (after demodulation), being derived from different speech, and having tones of different frequencies.

answers to tutorial questions

Q1 suppose the lowpass filter bandwidth is B Hz.. Let the input signal be $\cos\omega t$ and the local oscillator be $\cos \omega_0 t$. For simplicity we assume these are very much greater than $2.\pi$ B.

After taking the product of the high frequency signals, the difference frequency component, which is the only one likely to pass through the filter, is $\cos(\omega_i - \omega_0)t$.

Provided $|(\omega_i - \omega_0)| < 2.\pi$ B then the difference signal will pass through the lowpass filter.

Thus ω_i can be either above or below ω_0 for this condition to be met, and so it can lie in a window of width 2.B Hz.

Q2 let the oscillator be $\cos\omega t$, and the input signal; $\cos(\omega+\mu)t$ - this is above ω rad/s.

Although a single filter is shown at the output of the summer, this could be replaced by two similar lowpass filters, one each in the P and Q arms. Then the lowpass terms into the summer will be:

from the P arm ===> $\cos\mu t$

from the Q arm ===> $\sin\mu t$, which is changed to $-\cos\mu t$ by the 90^0 network

There is a finite output if these are *subtracted*.

Now suppose a signal appears at the input, lower in frequency that ω rad/s. For the same circuit conditions the lowpass terms into the summer will be:

from the P arm ===> $\cos\mu t$ since $\cos(\mu t) = \cos(-\mu t)$

from the Q arm ===> $-\sin\mu t$, which is changed to $\cos\mu t$ by the 90^0 network.

There is no output, since the summer is acting as a *subtractor*.

Q3 you can show the maximum is broad (ill-defined), but the minimum is narrow (well defined).

Q4 if there is interference on one sideband only, the true SSB receiver can demodulate the other sideband. The improvement in output signal-to-noise ratio can be large. If the interference is removed, there is a 3 dB advantage to be had if both sidebands are recovered in a true synchronous demodulator (double the amplitude, four times the signal power; but only twice the noise power).

Q5

Q6 the carriers need to be at 90 degrees if, and only if, the QPS is perfect, and there are no other (although small, and fixed) phase shifts elsewhere in the circuitry . To account for these small imperfections then the phase of the two carriers into the two MULTIPLIER modules needs to be adjustable.

Q7 no ! a 10 Hz error goes undetected for speech. With a 100 Hz error speech is still intelligible to the experienced operator. For larger errors an experienced operator can still communicate, but it will now depend on other factors, including the *sign* of the error. It is easier to listen to an upward shift of say 200 Hz than downwards, as the percentage change is significantly different. Try it !

Note that the tolerable SSB carrier offset is an *absolute* measure. Thus the higher the carrier frequency, the tighter the tolerance on the receiver local carrier.

The sampling theorem

TRUNKS

Supply speech, unless students have SPEECH modules.

A two-tone audio test signal (else from the student`s SPEECH modules ?).

answers to tutorial questions

Q1 it prevents any out-of-band noise, accompanying the message, from being aliased into the passband.

Q2 samples would have been taken at a different time. This will introduce a phase angle into each of the AC components of s(t), but *not* the DC term. Hence each of the components of the sampled signal spectrum will be changed in phase except the wanted message term, which arose form the produce with the DC term of s(t). So there will be no change to the recovered message.

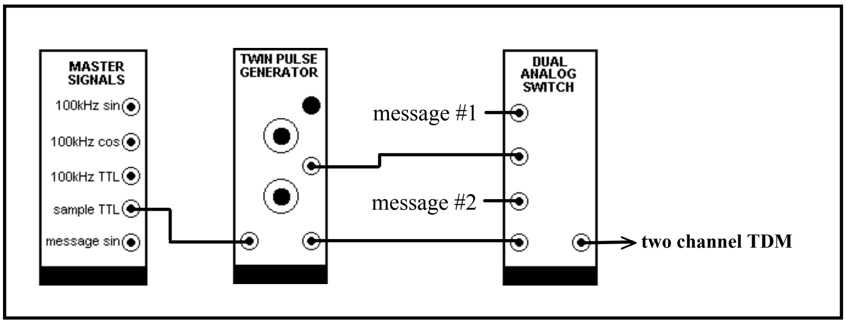
Q3 the switching function will have a faster pulse rate, so the unwanted spectral components will move further away from the edge of the reconstruction filter. Since the pulse width remains fixed, the ratio $\delta t/T$ is increasing (Figure 1-A of experiment); the amplitude of the recovered component will increase.

PAM & time division multiplex

TRUNKS

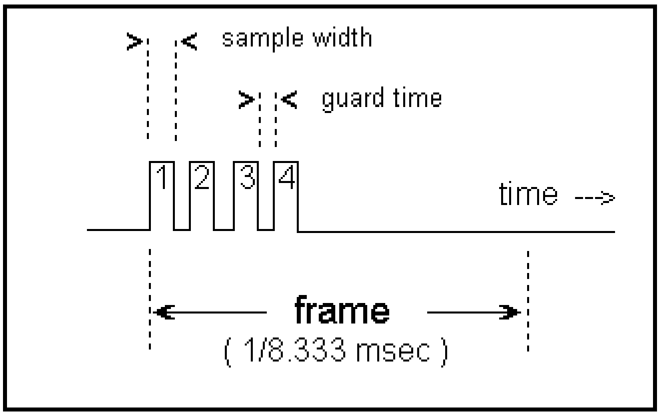
You must provide a four channel TDM signal.

Two channels of a TDM transmitter may be generated by the arrangement shown below. The sampling rate will be 8.333 kHz, so a frame occupies 120 µs. If you make the guard time between samples equal to the sample width, then the sample width will be 15 µs. But it could be made smaller (say 8 µs) and the active channels bunched to one end of the frame, to show more channels could be added (say 8 in total) if required.



two-channel TDM

This two channel TDM can be combined in an ADDER with another two-channel TDM to make a four channel system. Some phase shifting of the 8.333 kHz signal to one of the TWIN PULSE GENERATOR modules will be required. More actual channels are unnecessary, but more virtual channels can be implied by grouping these four down one end of the frame, implying more (inactive) channels.



sampling pulses to analog switches
(this could simulate an eight channel system)

Sources of suitable messages for each channel are discussed in the introduction to this ***Instructor`s Manual***.

answers to tutorial questions

Q1 if the switching pulse is initially narrower than the sample, then more of the sample will be passed to the reconstruction filter as it is widened, and the amplitude of the reconstructed message will at first increase (linearly) until the switching pulse becomes wider than the sample. From now on the recovered message amplitude will remain constant, but the signal-to-noise ratio will decrease. There will be an abrupt increase of crosstalk when the receiver starts accepting parts of the samples from the adjacent channel. The above remarks assume the switching pulse is located (in time) so as to straddle as much of the wanted sample as possible.

Q2 there is a minimum pulse width that is obtainable from the module, so this sets one limit. If this width could be narrowed further then eventually the rise and fall times would become a significant part of the pulse width, until eventually there would be no effective pulse at all.

At the receiver, as the sample widths fell, so would the signal-to-noise ratio of the recovered message, until eventually it would reach an unacceptable level.

Q3 see the answer to the previous question. The guard band allows for errors in the location of the samples, and of the switching pulse in the receiver. Practical pulses may have skirts which will overlap, so the guard band can help to reduce this source of crosstalk.

Q4 the linear sum of the messages from all channels.

Q5 see above patching diagram.

Power measurement

TRUNKS

Students should make their own test signals, so that they have control over all parameters. However, an extra audio sinewave would be useful, as would some speech.

answers to tutorial questions

Q1 1-a single tone. 2-two tones 3-DSBSC 4-AM 5-Armstrong`s signal 6-angle modulation 7-bandlimited speech is often quoted as having a peak-to-average power ratio of 14 dB. The entry in the rms column is based on this.

Q6 the meter is based on the AD637, which is described in data sheets available from Analog Devices.

INTRODUCTORY DIGITAL EXPERIMENTS

PRBS generation

TRUNKS

No signals required.

answers to tutorial questions

Q1 use your oscilloscope delayed time base facility, if available. Otherwise endeavour to insert a delay between the start-of-synch signal from the SEQUENCE GENERATOR and the oscilloscope ext. trig. input. The following methods are worth thinking about, although each proves inadequate at clock speed of 2 kHz (as used in this, and in many later experiments):

- use the TWIN PULSE GENERATOR, using the delayed pulse to trigger the oscilloscope. But the delay available at typical (2 kHz) clock speeds is insufficient.
- use the DIGITAL DELAY sub-system within an INTEGRATE & DUMP module. This provides an adjustable delay up to 1.5 ms, but this is inadequate.
- a TUNEABLE LPF (followed by the COMPARATOR in the UTILITIES module) will introduce a delay, variable with bandwidth. Once again, only a few clock pulses at most.

However: *for demonstration purposes*, the clock speed can be increased (even to 100 kHz) if you really must see the next ‘n’ bits, where ‘n’ is not too large. Then all methods are useful.

Q2 the rectangular pulse retains its approximate shape if the filter bandwidth is at least $1/(2\tau)$ Hz, where ‘τ’ is the width of the pulse in seconds.

Q3 *synchronized* if they use the same clock; *aligned* if they are the same sequence pattern, and both ‘start’ at the same time (where the actual ‘start’ occurs is a matter of definition).

Q4 the filter will introduce a time delay. This depends upon its order (‘size’), and bandwidth. The order is fixed, but the bandwidth, and so transmission delay, can be changed.

Q5 in general, for a longer sequence, more trials are necessary until alignment is achieved.

Q6 the counter counts transitions in one direction only. Each bit of the sequence starts at a clock HI and lasts for a clock period. If the sequence was a series of 0s and 1s, there would be half as many sequence transitions as clock transitions. Thus the counter would display half the clock frequency. If, in the sequence, some of the zeroes became ones, or vice versa, the sequence would contain less transitions; it cannot contain more. Thus the counter would indicate less than half the clock frequency. The pattern is such that it reads half of half the clock frequency.

Q7 if the TTL sequence is either 0 or 5 volt, it would generate half the power of a continuous 5 volt signal. Thus power in one ohm would be 12.5 watt. Thus rms value = $\sqrt{(12.5)}$ volt

Q8 the clock period is 1/2083 s, there are 2048 bits per sequence, one bit per clock period. So the delay is almost 1 second.

Eye patterns

TRUNKS

No signals required.

answers to tutorial questions

Q1 baseline wander is usually the result of an AC coupling. It is accentuated when the data has long sequences of identical values, which, with AC coupling, introduce a DC shift (wander). This upsets the decision device threshold. Appropriate line coding can transform the data so that such long sequences are avoided.

Q2 they are meant to be filters satisfying the same channel allocation requirements. This is more likely to be by means of a slotband definition than a passband width. See the definition of slotband (and other filter terminology) in Appendix A to Volume A1 of *Communications Systems modelling with TIMS*.

Q3 there are occasional traces which fall within the eye. They are more likely to be observed with a storage than a more conventional type of oscilloscope.

Q4 it exercises more data patterns

Q5 the transitions between top and bottom of the eye (at the corners of the eye) will be misplaced.

The noisy channel model

TRUNKS

No signals required

answers to tutorial questions

Q1 decibel is a normalised scale, which simplifies amplitude response comparisons. It allows responses of filters in cascade to be added. It preserves relative shapes for different scale ranges. Logarithmic frequency offers a useful compression of the scale; an octave covers the same width at all frequencies. For small frequency ranges (say less than an octave) it is less important (although it will show symmetry in the case, for example, of a second order bandpass filter). A disadvantage is that the absolute level of measurement is not shown explicitly, and this can be important information.

Q2 the sinewave will not appear distorted if the channel is narrow band, thus not passing distortion components (see the Part I experiment entitled *Amplifier overload*). The single voltmeter reading alone would not reveal the existence of overload. Two or more measurements should reveal the lack of linearity between input and output. In some cases two measurements would *not* be sufficient - in the case, for example, of 'fortuitous cancellation' (this was *not* discussed in the aforementioned experiment !).

Q3 assuming uniformly distributed noise power then the noise power would halve; thus the meter reading will be reduced by $1/\sqrt{2}$

Q4 $(5/2)(1/\sqrt{2})$ volt

Q5 a) $\sqrt{(25/2)}$ volt

b) $\sqrt{(25)} = 5$ volt

Q6 the square of the meter reading will give the power in one ohm. This does not lead to a linear conversion factor.

Q7 on noise alone the meter would read 0.458 volt rms. Thus SNR = 6.8 dB. Changes of less than 10% would be more difficult to read. Thus unless the SNR is say 10 dB or less the measurement of small changes to SNR can become difficult.

Q8 a) 3dB

b) reduced to $(1/\sqrt{2})$ of initial value.

Q9 using a sine wave as a test signal will show a delay (unless it is an exact multiple of 360^0), but cannot distinguish between delays of α and $(\alpha + n \times 360^0)$. Use the output from a SEQUENCE GENERATOR, clocked at well below B Hz (where B is the filter bandwidth). Then individual bits can be identified, and their transmission delay noted. There are other methods.

Signal detection with the DECISION MAKER

TRUNKS

No signals required.

answers to tutorial questions

Q1 the clocked/gated/strobed comparator makes only a *single* decision during a symbol period (be it correct or otherwise). This decision is made at what has been decided to be the optimum decision instant. An instantaneous comparator may detect a *multiplicity* of transitions within a symbol interval if there is a high level of 'ringing' (eg, due to a sharp cutoff filter with large phase distortion) or noise.

Q2 *timing jitter*: the decision instant has a random variation relative to the ideal. The result is that a decision may occur at a value that has a reduced margin relative to the noise.

baseline wander: see answer in previous experiment

Q3 as per text book.

Line coding

TRUNKS

No signals required.

duobinary encoding

There is no procedure given for investigating duobinary encoding, although this is incorporated in the line coding modules. You may wish to add something to the experiment to cover this topic.

answers to tutorial questions

Q1 the answer to this question could fill a book. Topics would cover:

- spectrum shaping (can minimize energy near DC especially for telephone line applications)
- enhancement of timing information for timing recovery at the receiver
- elimination of the DC component (to control baseline wander; eliminate waste of power)

and so on.

Q2 processing delay by both the LINE-CODE ENCODER and LINE-CODE DECODER modules.

Q3 reduction of baseline wander (qv). Transmission systems are often AC coupled, and so a DC component would be undesirable.

ASK - amplitude shift keying

TRUNKS

No signals required.

answers to tutorial questions

Q1 for synchronous demodulation once the carrier has been acquired then the bit clock is available by division.

Q2

Q3 the bandwidth of the ASK is twice that of the message - the data. Refer to the bandwidth of a DSB signal. It has nothing to do with the phasing of the message. So the bit clock/carrier phasing has no influence upon the amplitude spectrum of the ASK.

FSK - frequency shift keying

The BIT CLOCK REGEN module is essential.

In addition:

- an extra AUDIO OSCILLATOR or VCO: as a sinewave source in the range 1 to 5 kHz is convenient. This is the SPACE signal for transmission. But alternatively you can supply a suitable sine wave at TRUNKS.
- extra UTILITIES and TUNEABLE LPF modules are optional. Required if both channels of the demodulator are to exist simultaneously.

This experiment has no step-by-step Tasks. Only a collection of models to evaluate.

Preparation by students is important, especially with regard to the asynchronous demodulator. This uses the two BPF in the BIT CLOCK REGEN. module. Because the filters are required to be on different frequencies, one must be tuned by an external oscillator, the other to 2.083kHz (internal clock).

The BPF bandwidths are approx 2% of centre. Having one tone (arbitrarily called the MARK) at 2.083 kHz sets the bit rate many times below this. The divide-by-8 in the BIT CLOCK REGEN. module is used to lower an AUDIO OSCILLATOR to below 50 Hz.

Such low speeds make (standard) oscilloscope viewing difficult, but not impossible.

choice of two frequencies

If spectral analysis facilities are available much work can be done regarding spectrum control by suitable choice of the two signalling frequencies (as a function of bit rate). This would require the ability to generate such frequencies, a facility which TIMS currently cannot conveniently provide. In addition, some more appropriate filters might be required. This is perhaps getting too far from the intent of TIMS.

However, if users desire such a capability, a suitable module or modules could be developed for the purpose. Suggestions are welcome.

TRUNKS

See above.

1. a sine wave in the range 1 to 5 kHz - say 2.5 kHz.
2. an FSK signal - f_1 on 2.083 kHz and f_2 nearby (say 2.5 kHz).
3. perhaps a low frequency bit clock would be a help.

answers to tutorial questions

Q1 it is the sum of the two spectra.

Q2 carrier frequency not stable enough for communications applications; otherwise OK

Q3

Q4 the filter centre frequencies must be far enough apart for the filters to be able to separate the two tones. There are optimum spacings for minimum ISI. Signalling speed is limited by the inequality that requires

$$1/T < B$$

where B is the filter bandwidth, and T the period of the clock.

Q4 bandwidth would set a maximum bit rate. Minimum separation would be determined by the response shape. But, depending on filter characteristics, ISI can be minimized by certain critical spacings. Outside the scope of TIMS.

Q5 cost, complexity, required SNR

Q6 a TTL LOW is the green socket on the VARIABLE DC module.

- post-1998 TIMS have a TTL high available from the VARIABLE DC module.
- the DIGITAL DIVIDER module provides a TTL HI

Some digital modules (when *unclocked*) have their outputs at a TTL HIGH:

- the COMPARATOR (in the UTILITIES MODULE) with REF grounded
- DELTA MODULATION UTILITIES output (no input, or clock, connected)
- Q2 of the TWIN PULSE GENERATOR (Q1 is a TTL LOW)

Alternatively use a -ve output from the VARIABLE DC module, amplified and reversed in polarity through a BUFFER AMPLIFIER.

Q7 an SNR advantage.

BPSK - binary phase shift keying

TRUNKS

An 'unknown' BPSK near the carrier frequency used in the experiment could be transmitted. If you derive a stable clock from the system 100 kHz master then part of the requirement would be to report its frequency. Identification of the message sequence might be a problem (there would be no start-of-sequence information), although you might be able to generate a very short sequence.

answers to tutorial questions

Q1 agreed. It is transmitted by an analog transmitter, and it has a narrow (sub-octave) bandwidth.

Q2 some line codes may have DC components. AC coupling of the MULTIPLIER will inevitably introduce base line wander following repeated series of '1' or '0'.

Q3 not the amplitude spectrum, which is what is usually thought of, and usually seen on a spectrum analyser. This is a linearly modulated signal. The bandwidth is twice that of the message. The phase cannot alter the message bandwidth.

Q4 no ! the baseband filter will impart a certain shape to the message sequence spectrum. This shape will be reflected (in mirror image form) in either side of the BPSK spectrum, no matter what the carrier frequency or phase). This is a linearly modulated signal.

Q5 it removes the sum components from the multiplier output - around twice the carrier frequency - and should be wide enough to pass sufficient of the signal to allow the detector to regenerate accurately. The narrower the passband the less the noise, but the more the degradation of the signal. A compromise.

Q6 delay through the filter changes with bandwidth (area under the delay/frequency curve remains constant). Depending upon the order of the filter it can change multiples of p. So the phase between the received signal and local carrier could rotate a significant amount, and so the amplitude could go up (if not already at a maximum).

Q7 phase between the local carrier and the resultant of each pair of sidebands.. This is easy to show using phasors. It is not easy to measure directly (impossible ?), but very easy to optimise.

Signal constellations

Read the *Advanced Modules User Guide* !

To save time, set the following on-board jumpers:

- M-LEVEL ENCODER: J3 to NORMAL position.
- M-LEVEL DECODER: the range to HI (for an 8.333 kHz clock).

TRUNKS

No signals required.

answers to tutorial questions

Q1

Q2

Q3

Sampling with sample-and-hold

TRUNKS

None. But you may like to supply speech if SPEECH modules not available.

answers to tutorial questions

Q1 an engineering estimate is that one can see the presence of the smaller amplitude component when it is about 30 dB below the larger component (by CRO observation). Certainly a 40 dB ratio is *not* visible.

Q2 put the output of the reconstruction filter into an ADDER. Combine with the original message via a PHASE SHIFTER. Null the message; remainder will be *at least* 40 dB below - all that TIMS claims;   probably better than 50 dB below. Can`t see any distortion due to the S&H process.

Q3 the slot band is the sum of the passband width and the transition band. That is, from DC to the start of the stopband. Nyquist assumes a brick wall characteristic (zero transition band). So the re-written Nyquist criterion, assuming a bandlimited message, would say that the sampling rate should be twice the message bandwidth plus the transition bandwidth. Including the effects of practical message bandwidth limiting, the sampling rate would be twice the message bandwidth plus *twice* the transition bandwidth (assuming similar filters for bandlimiting and reconstruction).

The first of these engineering definitions is applicable when testing with a single sinewave (f_o), since the message appears to be brickwall filtered to f_o.

Q4 cancellation of the aperture effect with an equalizer; possible because the reconstruction filter introduces *linear* frequency distortion. See sinx/x correction.

PCM encoding

TRUNKS

No signals required.

oscilloscope triggering

Using the frame synch signal FS for oscilloscope triggering is useful for showing frames, but if the sweep circuits 'miss a beat' then adjacent frames will swap positions on the screen. Sometimes this is a nuisance. If the FS signal is divided-by-2 this cannot happen. Use the divide-by-two sub-system in the BIT CLOCK REGEN module.

WARNING

Please note that the maximum clock rate for this module is about 10 kHz. It is safe to operate at the clock rate of 8.333 kHz for which the module was optimized, but no faster.

answers to tutorial questions

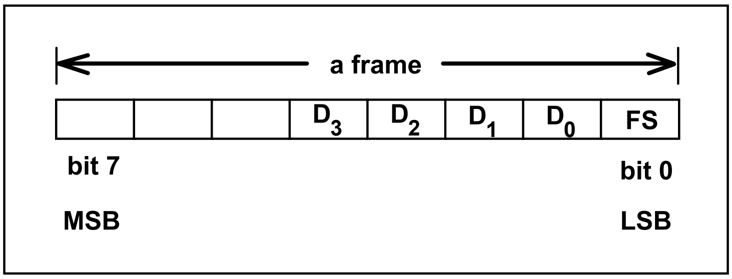
Q1 the clock rate to the module is 8.333 kHz. Frames are 8 clock bits wide. The input signal is sampled once per frame. So the sample rate is 8.333/8, or about 1 kHz. To satisfy the sampling theorem the message must be limited to frequencies below half the sampling rate, or less than 500 Hz.

Q2 the module is driven by an external clock. Samples of the input analog waveform are taken every eight clock bits, and coded (A to D) into either 4-bit or 7-bit words. These words are fitted into a time frame (together with a frame synch pulse FS) which is 8 clock bits wide.

The frame slots are numbered 0 to 7 as shown below.

For the case of a 4-bit word, its four bits fit into the frame as shown, D_0 being its least significant bit in slot #1. Slots 5, 6, and 7 are empty.

Slots 5, 6, and 7 are occupied when 7-bit code words are used.



Q3 the sequence of frames could be stored, and re-transmitted at a slower rate. This would be advantageous under poor SNR conditions, for example.

Q4 all samples of a DC message are identical. If the oscilloscope is synchronized to show one frame per sweep - to overlap consecutive sweeps - then the display will be 'stable'. Consecutive samples of an AC message will differ, and so when overlaid the display will not be stable.

Q5

Q6 if alternate frames of a PCM signal were omitted then it would appear that the input signal ('A', say) is being sampled at half the original rate. A second analog message ('B', say), sampled in the same manner, but with its frames timed to interlace with those of the first, could be added to it.

The message 'A' could be recovered by a decoder if the decoder examined only the 'A' frames; likewise for the 'B' message.

The TIMS PCM ENCODER has an operational mode (TDM) were alternate frames can be omitted; two such encoder modules can be set up to operate as master ('A', say) and slave ('b', say).

Because the sampling rate is effectively halved, as compared with the 'normal' usage, the allowable message bandwidth is also halved in this mode.

Q7 advantageous where the message is speech.

PCM decoding

TRUNKS

None. It would be advantageous to demonstrate the effects of companding with speech as the message. Unfortunately the processors used in the PCM modules restrict the system bandwidth to well below that of speech.

Oscilloscope Triggering

Using the frame synch signal FS for oscilloscope triggering is useful for showing frames, but if the sweep circuits 'miss a beat' then adjacent frames will swap positions on the screen. Sometimes this is a nuisance. If the FS signal is divided-by-2 this cannot happen. Use the divide-by-two sub-system in the BIT CLOCK REGEN module (or divide-by-4 in the LINE-CODE ENCODER).

WARNING

Please note that the maximum clock rate for this module is about 10 kHz. Thus it is safe to operate at the clock rate of 8.333 kHz.

PCM DECODER

For message reconstruction an LPF operating down to about 200 Hz is required. The TUNEABLE LPF will not tune as low as this. Version V2 of the PCM DECODER module has a built-in LPF for this purpose; so it is advantageous to supply this up-dated version if possible.

answers to tutorial questions

Q1 there will be a delay though a bandlimited channel. The phase adjustment acts as a delay adjustment to the stolen carrier to compensate.

Q2 the two waveforms, as seen on an oscilloscope, look, at a first glance, to be the same. The number of steps in each waveform is exactly the same, since sampling rate is the same. The difference is in the accuracy with which the amplitude of the steps represents the amplitude of the sampled message. The more accurate (7-bit), the less will be the distortion of the recovered message.

Q3

Q4

Q5 a) nothing. b) increase the number of quantizing levels. This requires a faster bit rate, and so more bandwidth, if the word rate is to remain the same. But see next Question

Q6 more quantizing levels - price to pay? To keep the same word rate the bit rate would need to increase. This means more bandwidth. BUT: in the present system the bit and word rate remains the same because of the way the frames are formatted. For both 4-bit and 7-bit coding schemes the words are all 8-bits wide. So the data rate is the same; the channel is not used efficiently.

Q7 the advantages show up when speech is the message. Intelligibility measurements are non-trivial, so you may or may not be convinced by a quick listening test.

Delta modulation

There are many parameters which can be varied in the delta modulator, and investigating the effects of all of them is time consuming and perhaps unnecessary.

If you have certain preferences you could instruct the students to concentrate on these, rather than following the experiment as written.

Note that not much attention was devoted to a change of clock speed.

You may prefer to insert a separate, non-inverting amplifier between the INTEGRATOR and the ADDER, rather than use the ADDER gain **g** to change the loop gain. A change of ADDER gain **g** alters the step size, but one cannot see this as the change occurs inside the ADDER.

TRUNKS

Speech is optional (either via TRUNKS or using student`s own SPEECH modules).

answers to tutorial questions

Q1 stable picture; message is a sub-multiple of the sample rate

Q2

Q3 highest message frequency determines the maximum message slope

Q4 $\sqrt{3}$

Q5 one bit per word.

Delta demodulation

TRUNKS

Speech is useful (either via TRUNKS or using student`s own SPEECH modules).

answers to tutorial questions

Q1 for a given message slope, and sample rate, there is a step size below which slope overload will occur. To avoid this the step size must be increased. When the message slope is low a small step size will result in smaller errors. Thus avoidance of slope overload requires large steps, which will increase the error in the regions of low slope (giving rise to what is referred to as granular noise).

Q2 this scheme allows the measurement of (signal + noise), and then noise alone. The method is described with reference to Figure 2 in the Experiment.

Q3

Q4 cancellation of more than one frequency component simultaneously would require more than a simple phase shifter.

Q5

Adaptive delta modulation

TRUNKS

Speech is optional (either via TRUNKS or using student`s own SPEECH modules).

Delta-sigma modulation

The operation of this modulator-demodulator system can be described in simple terms. But this will probably fail to reveal the reasons for its application in the CD player, where both cost and performance are paramount.

The experiment ‘works’, but it has been left to you, the instructor, to flesh out the experiment details to complement the level of your course work.

FURTHER AND ADVANCED ANALOG EXPERIMENTS

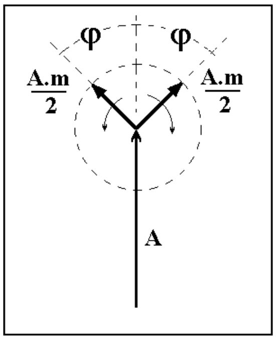
Amplitude modulation - method 2

TRUNKS

Speech is useful; perhaps students have their own SPEECH modules ?

answers to tutorial questions

Q1



The phasor diagram represents the AM signal $y(t) = A(1 + m.\cos\mu t)\cos\omega t$. The amplitude of the AM signal is the resultant of the three phasors. The resultant of the DSBSC is co-linear with the carrier phasor, since the two angles φ are equal.

The trough of the AM signal occurs when this sum is at a minimum. You can see the sum will be zero when φ = 180 deg, and m = 1

If the resultant of the DSBSC is not co-linear with the carrier but offset by an angle α (not zero), then the resultant phasor of the DSBSC can never combine with the carrier phasor to produce zero amplitude.

Q2 you cannot use a 'normal' commercial phase meter ! It would not know the phase of the *resultant* of the DSBSC signal, unless it had special features for accepting such signals. Phase meters typically compare zero-crossings of the two input signals, which need to be periodic. A DSBSC is seldom (ever ?) periodic. Convince yourself that this is so.

Q3 this is a philosophical question which you should consider at your leisure.

Q4 the TIMS FREQUENCY COUNTER counts the number of times the input signal passes through a pre-determined amplitude level in the same direction. This level is not zero; a finite amplitude level is chosen, to avoid corruption by noise (and false readings with no input).

It can be shown that, provided the depth of modulation of an AM signal is less than 100%, its zero crossings are uniformly spaced at the carrier rate.

Thus, provided the depth of modulation is not so high that the envelope troughs fall below the level at which the counter counts, it will read the carrier frequency.

Weaver`s SSB generator

TRUNKS

Speech is interesting; but perhaps students have their own SPEECH modules ?

the experiment

This experiment, as noted, requires an extra pair of MULTIPLIER modules, an extra TUNEABLE LPF, and an extra PHASE SHIFTER module.

The phase shifter at message frequencies is required to produce the inphase and quadrature carrier pair at 2.083 kHz. If you do not have a spare PHASE SHIFTER module these two can be produced with a QPS module, rather than a PHASE SHIFTER module. If the input to the QPS is the 2.083 kHz sinewave from the MASTER SIGNALS module, then the P and Q outputs are in phase quadrature. There is no phase trimming possible, but this can be an advantage, since it removes one degree of freedom in the setting up procedure !

The **disadvantage** is that the presence of the QPS module might make some students think they are modelling some sort of hybrid phase-shift method of SSB generation.

One of the points often omitted from text books is the fact that Weaver`s method, if out of alignment, places the unwanted products in the same band as those wanted; so they do not interfere with adjacent channels. The unwanted components, being in your channel, will seldom go unnoticed by you (as they might if they fell outside it).

answers to tutorial questions

An important feature to notice, when analyzing Weaver`s generator, is the fact that the *unwanted* components fall on top of (occupy the same spectral space as) the *wanted* components. Thus an imperfect generator degrades its own channel, and not an adjacent channel.

Weaver`s SSB demodulator

TRUNKS

A sinewave just above, and a sinewave just below 100kHz, would be useful as test signals. These could come from two VCO modules - say 99kHz and 102kHz. These would simulate a 1kHz message on a LSB, and a 2kHz message on an USB (of a 100kHz carrier).

If TRUNKS is not available, then supply each student with a second VCO to generate either of the above signals. The object is to demonstrate that when the demodulator is set up to receive a LSB then a 99kHz signal will produce an audio output (1kHz) while there will be no audio output with the 102kHz input signal.

answers to tutorial questions

Q1

Q2 in principle complete cancellation of a single unwanted sidefrequency can be achieved if we have control of both the amplitude and phase of the signals into the summer. For this to be achieved over the whole sideband then the two filters would need to be *identical,* else degree of cancellation of the unwanted sideband will vary with frequency.

For the single filter case the possibility of overload of the multiplier is increased, since both the difference *and* the sum frequencies will be present at its inputs.

Carrier acquisition and the PLL

TRUNKS

1. a nominal 50 kHz sinusoid. This avoids the need for students to have two VCO modules.
2. a DSBSC, based on a 50 kHz carrier, with a single tone message.
3. A DSB based on a 100 kHz carrier, derived from a single tone, with some (pilot) carrier added. This is conveniently made by using, as the message to a DSBSC generator, a tone plus a DC voltage, which have been combined in an ADDER. The DC voltage will provide the carrier term. Keep the ratio of the DC to AC amplitude small, so that the 'pilot' carrier will be relatively small. See Figure 1 of the experiment.
4. a DSBSC, based on a 50 kHz carrier, with a speech message. This is required for the final part of the experiment, which you may elect to demonstrate instead (see below)

Since only three TRUNKS channels are available the first signal will need to be removed towards the end of the experiment, and replaced by the fourth.

The VCO is described in the *TIMS User Manual*. Pertinent information for this experiment is as follows.

To obtain a 50 kHz output:

1. before plugging the VCO in, locate the on-board control RV8
2. toggle the on-board switch SW2 to FSK mode
3. plug in the VCO, leaving two free slots on the left for hand access to the board
4. connect a TTL HI to the DATA input
5. select HI on the front panel toggle switch
6. tune the on-board control RV8 (the front panel f_o control is inoperative in FSK mode) for 50 kHz (using the FREQUENCY COUNTER).

the experiment

You may elect to instruct students to miss the final part of the experiment, which requires them to have three MULTIPLIER modules. Instead, you could set up a demonstration for them to examine.

The loop filter for the PLL is in the UTILITIES module. This has a 3 dB point at around 3 kHz. It would be preferable if this were lower in frequency. There is such a loop filter in the BIT CLOCK REGEN module. If you have any such modules, please supply them to students.

Version 2 of the UTILITIES module offers a choice of corner points, via an on-board jumper.

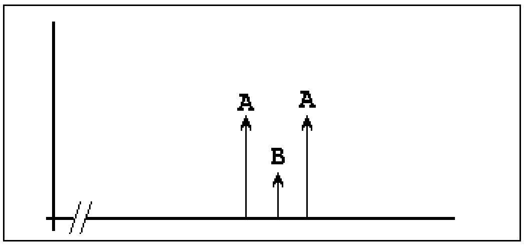
answers to tutorial questions

Q1 the BPF will not track the incoming signal, so its bandwidth needs to accommodate transmitter frequency drifts. It is harder to design, more complex to build, difficult to make tuneable.

The PLL has a fixed-frequency LPF, easy to design, the PLL is tuneable and will track. The PLL output is of constant amplitude; remains there if signal fades momentarily.

Q2 the signal has major and minor envelope peaks. Let the major peak-to-peak amplitude be P, and the minor Q.

Knowing the spectrum is as illustrated below (and given the carrier is in phase with the sidebands):



then it follows that (2A + B) = P and (2A - B) = Q

From Figure 1 P/Q ≅ 2.2/2.0

Thus A = 1.05, B = 0.1

Relate the carrier peak amplitude to the signal peak amplitude

$$carrier\,level = 20\log_{10}[\frac{B}{2A+B}]\,dB$$

This evaluates to about -26 dB.

Q3 the tuning of the PLL is by an oscillator - simple. To tune a BPF is far more difficult. The design of, and realization of, a BPF is more complex than that of a LPF.

Q4

Q5 the 'two signals' are a carrier on 100 kHz, and, in effect, a DSBSC. The DSBSC from TRUNKS can have undergone delays - phase shifts - different to those experienced by the system 100 kHz clock. So these two may no longer be in phase at a remote point.

Q6

Q7 the PLL will have constant amplitude; will be present during fades (although may drift off frequency if the fade is too deep); will track the signal if it drifts in frequency

Spectrum analysis - the WAVE ANALYSER

This experiment introduces students to a new module.

SPECTRUM UTILITIES module

This module is not in the TIMS BASIC MODULE SET. It uses a centre-reading moving coil meter movement, preceded by a buffer amplifier/filter.

The moving coil meter is indeed a mechanical filter, and will not respond to an AC signal above say 10 Hz; certainly not to 100 Hz. It is this that gives the analyser its selectivity.

The measurement of the amplitude of spectral component takes a little skill in tuning the VCO to within 1 Hz or so of the unknown component, and then estimating the peak reading of the meter as it oscillates slowly about the centre zero. You might agree that to expect more than a 10% reading accuracy would be unreasonable.

It would seem natural, then, to include some electronics to capture and hold the peak meter reading. This facility has been included in the module`s circuitry. *But it must be used with caution, and with the full understanding of the likely mis-interpretations*.

The sample-and-hold circuit connected to the signal across the meter has no frequency selective properties - it will hold what ever voltage is present.

If the peak meter reading without the sample-and-hold, and the sample-and-hold reading, are to agree, there must be only one component of significance across the meter.

This condition can be approximated provided:

1. spectral components are separated by at least say 500 Hz
2. the amplitude ratio of adjacent components does not exceed, say, 100:1

The second requirement is included to cater for the case where a small component is being measured in the presence of a larger adjacent one.

To aid the mechanical filter the buffer amplifier has a lowpass characteristic, with a corner at about 10 Hz, and beyond this an attenuation rising at 24 dB/octave.

TRUNKS

Unknown signals are required for spectral identification. You can choose those signals which you think would be appropriate for your students. Put them in both the audio and 100 kHz region.

Use an ADDER to combine signals from the VCO, AUDIO OSCILLATOR, and so on.

Suggested signals could be

- two-tone audio
- a low audio frequency square wave
- 100 kHz DSBSC *plus* an output from a VCO
- etc etc !

answers to tutorial questions

Q1 the analyser has a window 2 Hz wide. This is tuned up and down the input spectrum. The point of the question is to show that the bandwidth of the instrument is the *same* as that of the filter.

Q2 the instrument is of the same form as a product demodulator (for DSB signals). It looks out over a width B *either side* of the centre frequency. So its bandwidth is *twice* that of the associated LPF, or 2 Hz.

Q3 before the signal reaches the analyser it is distorted - new components are generated. Where are they ?

Let:

$$DSBSC = E.\cos\mu t\cos\omega t$$

With this as the input, the non-linear amplifier output will be:

$$e_{out} = a_1 E\cos\mu t\cos\omega t + a_3 E^3\cos^3\mu t\cos^3\omega t$$

Now we are only interested in components near ω, so it is not necessary to carry out the complete trigonometrical expansion.

Observe that the original signal is present in the output, scaled by a_1.

The distortion components represented by $\cos^3\mu t$ form the message to a DSBSC on both ω and 3ω (which we know will come from the $\cos^3\omega t$ term). We only need consider the first of these DSBSC. Thus components near ω are:

$$e_{out} = \frac{a_1 E}{2}[\cos(\omega+\mu)t + \cos(\omega-\mu)t] + \frac{3a_3E^3}{4}(\cos^3 \mu t)\cos\omega t$$

Expanding this further the spectrum near ω is:

frequency	***amplitude***
ω - μ	$\frac{a_1E}{2} + \frac{9a_3E^3}{32}$
ω + μ	$\frac{a_1E}{2} + \frac{9a_3E^3}{32}$
ω - 3μ	$\frac{3a_3E^3}{32}$
ω + 3μ	$\frac{3a_3E^3}{32}$

See the experiment entitled *Amplifier overload*.

Q4 only turn on HOLD when within a few Hz of the term being sought. Keep turning the HOLD facility off to ensure the correct signal is being held.

Amplifier overload

TRUNKS

No signals specified.

answers to tutorial questions

Q1 the tones are often kept close to ensure that the intermodulation products won`t be spread apart too far. It also makes the appearance in the time domain more like a DSBSC. However there are no hard and fast rules. Other applications make the two tones far apart, and not necessarily of equal amplitude.

Q2 the DSBSC method of making a two-tone signal is very convenient. It enables them to be moved across the spectrum by varying the frequency of the higher-frequency term, and their separation by varying the low frequency term. At the same time it provides a signal for oscilloscope synchronization (the low frequency term).

But the weak link is the MULTIPLIER, which is likely to be the cause of intermodulation distortion. The test signal will thus be impure before it even reaches the circuit under test.

Combining two tones with an ADDER is a much more linear process.

Q3 adjustment of the relative amplitudes of the two tones to equality is a simple matter of obtaining a symmetrical DSBSC-like display on the oscilloscope. Only when they are of equal amplitude can they add to zero and produce a null in the envelope

Q4 unstated in the question, and possibly in the advertisement, is the accompanying amount of distortion at the quoted power output. Presumably the '60 watt' amplifier will generate more distortion when delivering an output of 60 watt than will the '50 watt' amplifier when delivering 50 watt. One business may sell for the HI FI market, and the other PUBLIC ADDRESS applications. The requirements of these applications are quite different.

Frequency division multiplex

TRUNKS

An FDM signal is required.

You can model the generator with TIMS, and connect the output to TRUNKS.

The experiment says the signal can be recorded on tape. So it will be more convincing if you do this, and then play the tape back during the experiment. It will certainly save time for future presentations of the experiment.

Reference to the experiment will remind you that two schemes are suggested. There is little difference in the difficulty of making either recording (your responsibility), but the method of demodulation (student responsibility) is significantly different.

The experiment has been written on the assumption the simpler arrangement of Scheme 2 will be used. Although it is not in principle necessary to build two demodulators it may be more convincing to do so, and so the notes recommend this.

FDM recording

For the recording you will need a good-quality audio tape recorder. Bandwidth per channel should be at least 16 kHz, and signal-to-noise ratio at least 60 dB.

In addition at least two general-purpose cassette recorders are required (or four if the more ambitious scheme is attempted), together with cassettes of different passages of band-limited (300-3000 Hz) recorded speech. The preparation of these has been described earlier.

method 1 - scheme 1

It is possible to record all FDM channels simultaneously, on a single track, according to the scheme illustrated in Figure 1 below. Notice that an *upper* sideband is required from each SSB generator

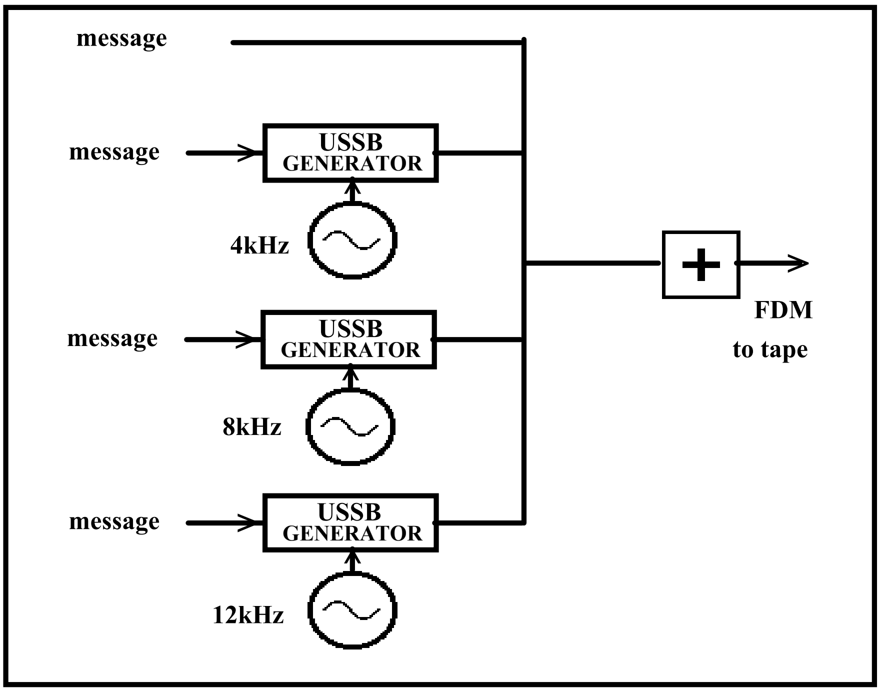


Figure 1

Whilst the above scheme looks elaborate, it is well worth the trouble to get the complete FDM signal onto tape. From then on setting up of the experiment is greatly simplified.

method 2 - scheme 1

A less elaborate scheme for preparing the recording is suggested in Figures 2 and 3. This builds up the recording in two stages, on two different tracks. The two tracks can then be combined onto a single track of a separate high-quality recorder, or the two tracks can be combined in an external ADDER at the time of transmission via TRUNKS.

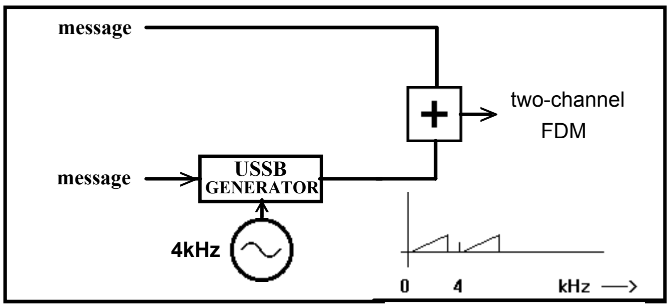


Figure 2

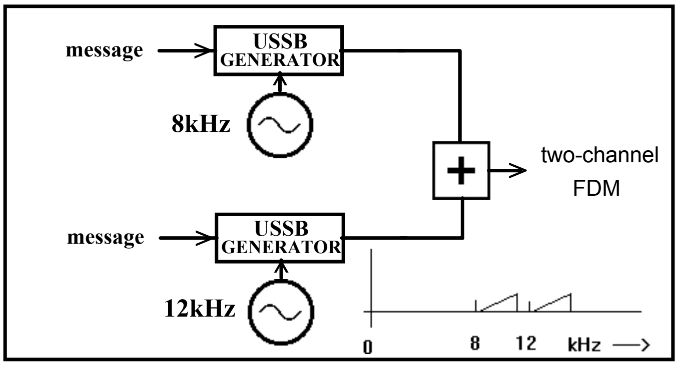


Figure 3

method 1 - scheme 2

It is possible to record all FDM channels simultaneously, on a single track, according to the scheme illustrated in Figure 4 below. Notice that one generator produces a *lower* sideband, and the other an *upper* sideband.

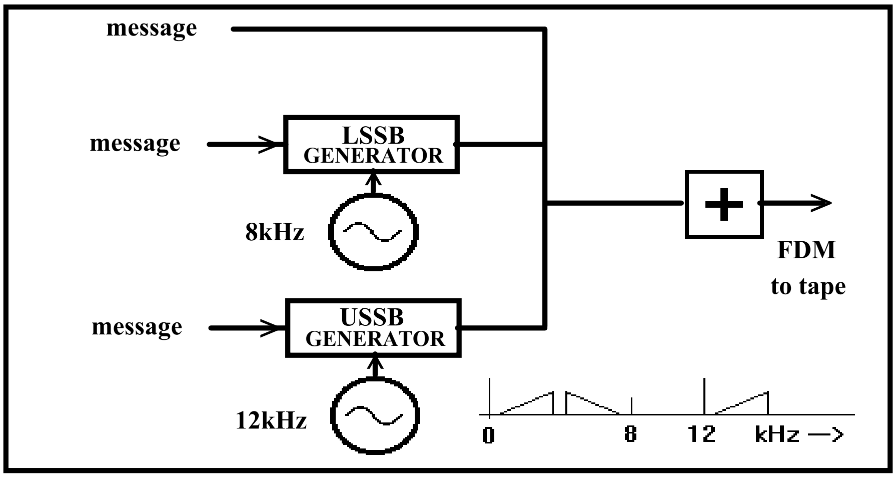


Figure 4

method 2 - scheme 2

A less elaborate scheme for preparing the recording is suggested in Figures 5 and 6. This builds up the recording in two stages, on two different tracks. The two tracks can then be combined onto a single track of a separate high-quality recorder, or the two tracks can be combined in an external ADDER at the time of transmission via TRUNKS.

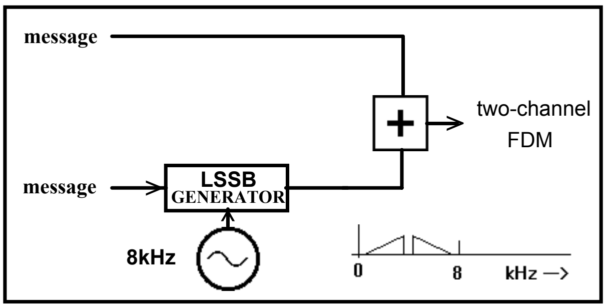


Figure 5

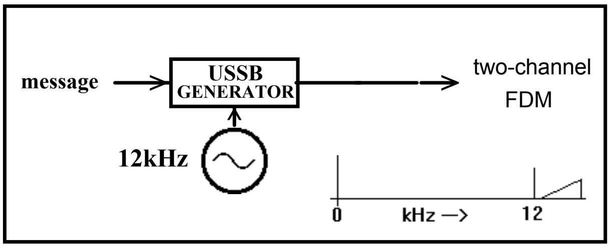


Figure 6

other recording schemes

You may think up other schemes for preparing the FDM signal.

Be aware that the isolation between tape recorder tracks is often not great with 'stereo' type cassette recorders. Although this should not be a problem with the two-track scheme suggested above, you should be aware of it if problems arise.

recording tips

- take care with setting up for the recording. Avoid earth loops and other sources of mains hum and noise. Check the final result by *doing the experiment* yourself ! Unnecessary noise and extraneous signals must be minimized so the student will not be distracted.
- at all times during the preparation of the FDM signal, and later when trunking it, avoid any form of overload and thus crosstalk.
- make sure all messages are different, and bandlimited 300 to 3000 Hz. Whilst music makes a change from speech, it tends to have occasional high level passages which can overload any part of the system and thus cause crosstalk.
- if the tape recorder has any noise minimizing schemes, *disable them*. These often assume certain properties of the signal to be recorded (based on the assumption it will be music), which the FDM signal does not have.
- determine the amplitude V_{pk} of a sine wave into the recorder which just overloads the tape recorder. Do not rely on any built-in record level metering, but play back the tape and observe the waveform of the output. When making recordings ensure that the peak signal level into the recorder is at least half V_{pk} to avoid non-linearities and thus crosstalk between channels.
- ensure all SSB generators are well balanced, especially for carrier leaks.
- ensure the peak amplitudes of each channel into the ADDER (of Figure 1, 2, or 3) are equal.
- when combining the schemes of Figure 2 and 3 make sure peak amplitudes into the combining ADDER are equal.
- make sure the peak amplitude of the FDM into the TRUNKS system does not exceed the TIMS ANALOG REFERENCE LEVEL.

variation of the experiment

If the amplitude of the FDM from TRUNKS is too large, it will overload the demodulator. This is investigated in the experiment, circa Task #T7. It will be the MULTIPLIER which will overload.

You may prefer to simulate a channel overloading by interposing the COMPARATOR from the UTILITIES module, as was used in the ***Amplifier Overload*** experiment. This has a more controllable non-linear characteristic.

answers to tutorial questions

Q1

Q2 Yes ! The DSBSC demodulator will examine a window 3 Hz either side of its local oscillator frequency. In each case there is a wanted channel on one side, and nothing on the other. The carrier frequencies used would be 8 and 16 kHz

Q3 in principle the true SSB demodulator will have a signal-to-noise advantage, since it will not add the noise from the unused sideband.

Q4 Yes ! But a DSBSC demodulator scheme needs a synchronous carrier. Carrier acquisition would be a problem, but pilot carriers could be sent. Alternatively a master carrier could be recorded on the tape (say 16 kHz), from which the others could be derived by division. Carrier acquisition circuitry would need to track the carriers - they would move due to the non-constant speed of the tape recorder, and/or phase jitter. In principle this sounds easy - in pratice it will present problems in implementation.

An (asynchronous) SSB demodulator could recover the individual channels - this would be adequate for speech. This is an acceptable solution both in principle and in practice.

Q5 the guard band takes into account the finite width of the transition band of the channel filters, and to a lesser extent the frequency stability of the demodulating carrier oscillators.

Q6 how does one measure the peak amplitude of a speech channel ? How does one measure the peak amplitude of an FDM signal ? The oscilloscope, as a readily available and affordable instrument, is very useful. Specialised instruments also exist, but are expensive. If the peaks occur only occasionally could not one operate at a higher level than the peak level ?

Phase division multiplex

Whilst PDM does not find much application as a method of multiplexing speech in the 1990s, the principle involved is very important. This experiment is meant to illustrate the principle in a strictly analog environment. A later experiment (eg., QASK) shows a common application in digital communications.

TRUNKS

For this experiment you must send a PDM signal via the TRUNKS.

Each TIMS 301 has a copy of the TIMS 100 kHz clock, for use as the stolen carrier.

Make one message a single tone, and the other speech. Students can then assess the relative merits of nulling on a single tone, and on speech. It is easy to measure the depth of a null for a tone, but less so for speech. If the null of the tone is better than say 40 dB (as it should be) then it could end up below the system noise, and one can then only give an estimate of the null; eg., 'better than 40 dB'.

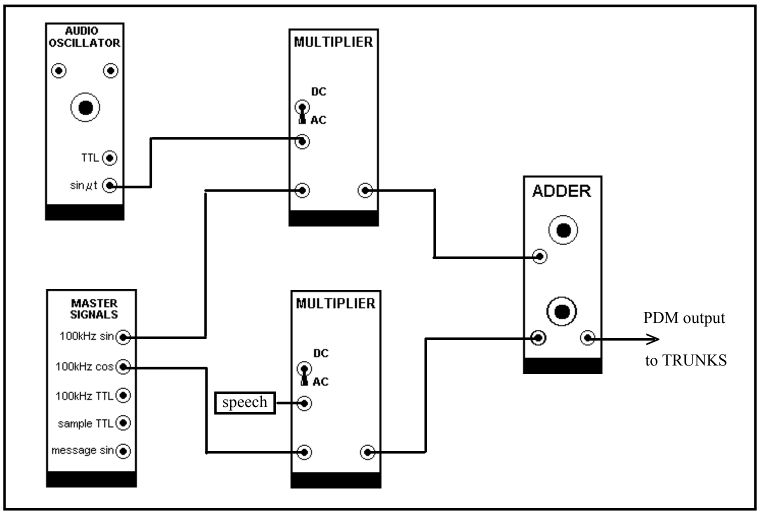
Take care not to overload the MULTIPLIERS, else non-linearity will spoil performance by introducing distortion components which cannot all be nulled at once at the receiver.

Do not allow the TRUNKS signal level to be such that the received signals at each TIMS 301 are above TIMS ANALOG REFERENCE LEVEL, otherwise over-load at the receiver will degrade the achievable depth of null. A BUFFER AMPLIFIER at the receiver could, of course, be used to adjust the amplitude from TRUNKS.

There is *no need* to trim the two carriers to exact phase quadrature. The *sin* and *cos* outputs from the MASTER SIGNALS module are quite close enough to quadrature.

You can note the frequency of the AUDIO OSCILLATOR, and make its measurement part of the experiment requirements. For more mature students this could be a two-tone audio signal, the measurement of the frequency of which requires a little more ingenuity.

patching diagram



PDM Generator

answers to tutorial questions

Q1 assume all noise output is independent of the local carrier phase. Thus we are concerned with the wanted signal amplitude only. The system is adjusted to null the output from the unwanted channel. The amplitude of the signal from the other channel will be proportional to the cosine of the phase angle between its resultant and the carrier.

At best the angle will be zero, and the multiplying factor will be cos(0) = 1. When off by 45 degrees the factor will be cos(45) = 0.707

Thus the amplitude ratio will be 0.707, and the power ratio will be 20 $\log_{10}(0.707)$ which represents a degradation of 3 dB.

Q2 the system depends on nulling the unwanted channel; what ever remains is declared as wanted. Thus a three channel system would have to null two channels at once. This is not possible.

Q3 consider each channel to carry a single tone of the same amplitude. Assume the two channels are at 90 degrees at the transmitter. From the answer to Q2 above, show that the crosstalk ratio would be

$$\text{ratio unwanted to wanted} = \sin(\alpha)/\cos(\alpha) \text{ or approximately } \sin(\alpha)$$

Thus a 1 degree error would make the crosstalk power ratio approximately 35 dB

Q4 the alternative arrangement as suggested in this question enables each channel to be adjusted independently, whereas in the arrangement of Figure 2 the phase shifters will be interactive. So which arrangement is preferable ?

What if the phasing was altered at the transmitter. How would the two compare ?

Analysis of the FM spectrum

answers to tutorial questions

Q1 perhaps, but the amplitude spectrum is missing the sign (phase) of the 'Bessel coefficients', as they are often called. So it is not a straightforward exercise. Some trial-and-error might help.

Q2 the amplitude E is a peak value, so:

$$power = \frac{[\frac{E}{\sqrt{2}}]^2}{50}\ Watt.$$

This is true for *any value* of β, but is obvious for β = 0.

Should you be tempted to obtain the power by summing the contributions from individual components don`t forget there is a single term at 'carrier' frequency, but the others come in pairs.

Q3 in considering the power in an angle modulated spectrum one uses the fact that it is independent of the degree of modulation. The power in the *unmodulated* carrier (β = 0) is taken as a reference. When power *ratios* are involved there is no need to convert amplitudes to rms values. Neither is there a need to include the amplitude `E` [as in eqn.(7)].

To answer the question one refers to Bessel tables, and takes the reference as `1`. Then, from the column for the appropriate value of β one adds the squares of the entries until the sum reaches the desired amount (ie., 0.9 in the present case).

The precaution necessary is that the carrier contribution is included *once*, whereas subsequent entries are included *twice* (to account for upper and lower sidefrequency pairs).

So, for β = 1, the entries are:

side-frequency	*relative amplitude*	*relative power*	*cumulative sum of powers*	*% of reference power*
0	.7652	.5855	.5855	58%
1	.44005	.1936	.9727	97.27%
2	.1149	.0132	.9991	99.1%
3	.01956	.00038		
4	.002477			

Thus for 95% of the total power only the first pair of sidefrequencies is required.

For β = 5, the entries are:

side-frequency	*relative amplitude*	*relative power*	*cumulative sum of powers*	*% of reference power*
0	.1776	.03154	.0315	3.15
1	.32758	.21462	.24616	24.61
2	.04657	.004263	.2504	25.04
3	.36483	.2662	.5166	51.66
4	.39123	.3061	.8227	82.27
5	.26114	.13634	.9591	95.91
6	.13105	.0343	.9934	99.34
7	.05338	.0057	.9991	99.91

Thus for 95% of the total power a bandwidth to include the first 5 pairs of sidefrequencies is required.

Q4 from the sidefrequency locations the message frequency is 2 kHz.

To find the power it is necessary to start from the carrier, and obtain a cumulative sum of the power in each sidefrequency until all 'significant' components have been included. This can be decided by waiting until the cumulative power approaches a limit.

	freq MHz	*volts*	*relative power*	*cumulative power*	*denormalized amplitudes*
0	100±0.0000	1.7461	3.0490	3.0490	0.2920
1	100±0.0020	0.4101	0.3364	3.3854	0.0686
2	100±0.0040	1.6330	5.3334	8.7188	-0.2731
3	100±0.0060	1.3111	3.4379	12.1567	-0.2192
4	100±0.0080	0.5480	0.6005	12.7573	0.0916
5	100±0.0100	1.9157	7.3401	20.0974	0.3204
6	100±0.0120	2.0944	8.7733	28.8707	0.3502
7	100±0.0140	1.5509	4.8106	33.6813	0.2593
8	100±0.0160	0.9004	1.6215	35.3028	0.1506
9	100±0.0180	0.4362	0.3806	35.6834	0.0729
10	100±0.0200	0.1826	0.0667	35.7502	0.0305
11	100±0.0220	0.0676	0.0091	35.7593	0.0113
12	100±0.0240	0.0225	0.0010	35.7603	0.0038
13	100±0.0260	0.0068	0.0001	35.7604	0.0011
14	100±0.0280	0.0019	0.0000	35.7604	0.0003
15	100±0.0300	0.0005	0.0000	35.7604	0.0001
16	100±0.0320	0.0001	0.0000	35.7604	0.0000
17	100±0.0340	0.0000	0.0000	35.7604	0.0000
18	100±0.0360	0.0000	0.0000	35.7604	0.0000

The relative power is entered as the square of individual components. This should be corrected by a factor which incorporates the load resistance, and takes into account the rms value of the amplitude. But for the present purposes this is unnecessary.

The amplitude E is equal to the square root of the cumulative power, so

$$E = 5.98 \text{ volts}$$

Using this to denormalize we get the final column of entries, which are the denormalized amplitudes, equivalent to the Bessel coefficients associated with each spectral component.

The formula quoted in Q1 can be used to determine beta, with the proviso that we don`t know the signs (although it is probable that they are positive for the first three terms) . If you try a few successive groups of three adjacent components you will find that most cases yield:

$$\beta = 7.25$$

If the signal is PM then this is the peak phase deviation.

For an FM signal this is $\Delta f/\mu$. Since $\mu = 2.\pi.f$, and f = 2 kHz, then

$$\Delta f = 91.1 \text{ kHz}$$

By examining the final column from the bottom up, until an amplitude exceeding 0.05 is reached (5% of 1, the amplitude of the unmodulated carrier), the number of sidefrequency terms satisfying the 5% significant criterion is seen to be 9. Thus, on that criterion:

bandwidth = (2x9xmessage freq) = 36 kHz

Q5 this represents a 180 deg phase change, not detected by the usual spectrum analyser. Such a phase shift on *one* component of a DSBSC pair of components has the effect of moving the resultant of the DSBSC by ninety degrees (quadrature). But when effective on both the net result is to reverse the sign (phase) of their resultant. Remember:

$$J_{-n}(\beta) = (-1)^n J_n(\beta)$$

Q6 the spectral lines will be spaced 1 kHz apart, with relative amplitudes obtained from the Bessel tables for β = 1, 5, and 10 respectively.

Q7 the number of sideband pairs required are respectively 3, 7, and 13. Thus the bandwidths are 15, 35, and 65 kHz respectively (based on a 2.5 kHz message). The bandwidth of a PM signal does increase linearly with message frequency, for a fixed peak phase deviation. But here there is a change of peak phase deviation, with no change of message frequency. The relationship between bandwidth (number of significant sideband pairs) and peak phase deviation is not a linear one.

Q8 for the 2.5 kHz message β = 20/2.5 = 8. On the 1% significant sideband criterion the number of sidebands required is 11, and the bandwidth = 55 kHz.

With an increase of message frequency by a factor of 4, β reduces to 2, the sidebands spread out, but less are needed. On the 1% criterion the number required is now changed to 4, so the new bandwidth = 80 kHz.

Q no change

Introduction to FM using a VCO

TRUNKS

No signals required

comments

For the two-tone message combine an AUDIO OSCILLATOR and the 2 kHz message from MASTER SIGNALS module in an ADDER. Use a BUFFER at the ADDER output to vary the two-tone amplitude.

answers to tutorial questions

Q1 outside of a communications environment, for example. For low power, short range situations such as non-critical remote control, telemetry.

Q2 sidebands will be spaced 1 kHz apart, either side of the central 100 kHz carrier. The relative amplitudes are calculated from Bessel tables, knowing $\beta = (5/1) = 5$

Thus:

#	*freq.*	*rel. ampl*
0	ω	0.1776
1	$\omega\pm\mu$	0.3276
2	$\omega\pm2\mu$	0.0466
3	$\omega\pm3\mu$	0.3648
4	$\omega\pm4\mu$	0.3912
5	$\omega\pm5\mu$	0.2611
6	$\omega\pm6\mu$	0.1311
7	$\omega\pm7\mu$	0.0534

Q3 sideband are spaced apart by 1 kHz. Relative to the amplitude of the unmodulated carrier (1), on the 10% significant sideband criterion 6 pairs of sidebands would be included, so B = 12 kHz.

Q4 the counter counts level crossings in a given time slot. Since the zeros are not evenly spaced in time, neither will be the count made by the counter. However, the count will depend on ratio of carrier to message frequency, message frequency, magnitude of the frequency deviation, as well as the period of the count. Try it !

Q5

Q6 a) for n = 1 the first zero for β between 3.8 and 3.9, the second between 7 and 8

b) for n = 3 there is a zero for β between 6.3 and 6.4

FM and the synchronous demodulator

TRUNKS

An FM signal derived from a single tone, based on a 100 kHz carrier (Armstrong). See the experiment entitled ***FM Deviation Multiplication*** for a suggested method of generation. Note that you will need an FM UTILITIES and a 100 kHz CHANNEL FILTERS module. The latter should be version 2 or later (V2 marked on circuit board). Filters shipped pre-June 1993 were V1, but this was not marked on the board.

The message tone should lie somewhere in the range say 500 to 1000 Hz. This is high enough to simplify resolution with the WAVE ANALYSER; not so high as to exceed the bandwidth of the filters you will need to use with the multiplier stages following the Armstrong modulator.

Use a frequency deviation to produce at least a few sidebands of significance. Ensure there is a component at carrier frequency so that the demodulator will have a DC output.

The aim of the experiment is principally to introduce the indirect (baseband) method of spectrum measurement.

answers to tutorial questions

Q1 since the message will be in the 1 kHz region, a sweep speed of about 1 ms/cm would be suitable for showing up a 1 kHz envelope.

Q2 the effect upon the amplitude of the wanted components would be slight, since it would be the difference between being scaled by sin(85), say, rather than by sin(90).

However, the components which should have been reduced to zero will now have small, but finite, amplitudes, being scaled by cos(85) rather than cos(90).

Q3 although the filter output would be at message frequency, there would not be a linear relationship between its amplitude and that of the message back at the transmitter. Worse, since the single tone message could lie anywhere in the audio range (say 300 to 3000 Hz) a filter suitable for 300 Hz would be useless for 3000 Hz (and vice versa).

Also, this is a special case of a single tone message; for a more complex message the question is almost meaningless.

Q4 this would have been apparent when first odd, then even, harmonics of the message were identified with the SPECTRUM ANALYSER.

Q5

Q6 first draw the amplitude spectrum of the baseband signal. These amplitudes will be the maximum attainable in each case. The corresponding 100 kHz (relative) amplitude spectrum is obtained by remembering that each of these AC components was measured by summing the contributions of *two* sidefrequency components around 100 kHz. The DC component (which you were responsible for by ensuring a significant component at carrier frequency) is derived directly from the single carrier term, and so does *not* need to be halved in amplitude. The amplitude spectrum, of course, does not show absolute phase, nor recognise the relative phases between components..

Q7 the power in the FM signal is proportional to its amplitude squared. It is also proportional to the sum of the squares of all (significant) components. Knowing the latter (from the amplitude spectrum) the former, E, can be found.

The amplitude of the n^{th} sideband from the carrier (n = 0) is known, and is $E.J_n(\beta)$.

Thus can be found $J_n(\beta)$? Not quite, since many values of β will satisfy the relationship. But by trying several values of n, and using a little ingenuity, the value of β can be isolated.

Q8 the DC term arose from translation of the *single* component at carrier frequency. All other terms arose from the translation of *two* components, which, after translation, fall on top of each other and reinforce to give twice the amplitude which would have been due to one of them alone.

Armstrong`s phase modulator

answers to tutorial questions

Q1 expand

$$
\begin{aligned}
y(t) &= a(t).\cos[\omega(t) + \phi(t)] \\
&= a(t).\cos\phi(t).\cos\omega(t) + a(t).\sin\phi(t).\sin\omega(t) \\
&= P.\cos\omega(t) + Q.\sin\omega(t) \\
a(t) &= \sqrt{[P^2 + Q^2]} \\
\phi(t) &= \tan^{-1} [Q/P]
\end{aligned}
$$

$$
\begin{aligned}
\text{thus for eqn(3)} \quad P & \quad E \\
\text{and } Q &= E.m.\sin\mu t \\
\text{so} \quad \text{envelope} &= E.\sqrt{[\,1 + m^2.\sin^2\mu t\,]} \\
\text{and} \quad \text{phase} &= \tan^{-1} (m.\sin\mu t)
\end{aligned}
$$

Each of the last two expression can be expanded as harmonic series in terms of μ. See appendix B for some ideas.

Q2 no ! It can typically only measure the phase between two sinusoidal components. For the DSBSC, it is the phase of the *resultant* which is required.

Q3 peak-to-peak = $\sqrt{(4^2 + 5^2)}$ volt; trough-to-trough = 5 volt

Q4 the phase adjustment is made when the amplitude ratio of DSBSC and carrier is unity. Call the carrier amplitude, and the DSBSC maximum resultant amplitude, E.

$$P = \sqrt{[(E + E\sin\alpha)^2 + (E\cos\alpha)^2]}$$

As the DSBSC resultant goes through its two maximum values twice per message cycle, the resultant amplitudes are P and Q, where:

$$P = \sqrt{(E + E\sin\alpha)^2 + (E\cos\alpha)^2}$$

$$Q = \sqrt{[(E - E\sin\alpha)^2 + (E\cos\alpha)^2]}$$

When $\alpha = 0$ these two are equal, but otherwise not.

The amplitude ratio is

$$\frac{P}{Q} = \frac{1 + \sin\alpha}{1 - \sin\alpha}$$

On an oscilloscope one could surely see a 5% amplitude difference ? In this case, solving for P/Q = 1.05 gives $\alpha = 1.4$ degrees

Q5 let the Armstrong signal be defined as:

$$y(t) = \sin\omega t + k\cos(\omega t + \alpha)\cos\mu t$$

where α is the error from quadrature, and the DSBSC to carrier amplitude ratio = k

Expand this into the in-phase and quadrature form and get:

$$y(t) = (1 + k\sin\alpha\cos\mu t)\sin\omega t + (k\cos\alpha\cos\mu t + k\sin\alpha\cos\mu t)\cos\omega t$$

which leads to

$$\Delta\phi = \frac{k\cos\mu t(\cos\alpha + \sin\alpha)}{1 + k\sin\alpha\cos\mu t}$$

As a first approximation let $\cos\alpha = 1$. Then

$$\Delta\phi = k\cos\mu t\ [\frac{1}{1 + k\sin\alpha\cos\mu t}]$$

The term in square brackets can be expanded by the binomial expansion to give the approximate result:

$$\Delta\phi \cong k\cos\mu t\ [1 - k\sin\alpha\cos\mu t + (k\sin\alpha\cos\mu t)^2 \ldots]$$

Further approximation (small α) leads to:

$$\Delta\phi \cong k\cos\mu t\ [1 - k\alpha\cos\mu t + (k\alpha\cos\mu t)^2 \ldots]$$

The first term represents the wanted term, the remainder unwanted.

Notice that, for $\alpha \neq 0$, odd and even harmonic distortion will be introduced (which is otherwise zero).

Q6 from the in-phase and quadrature form defined in the answer to the previous question the Armstrong envelope is:

$$a(t) = \sqrt{[(1 + k\sin\alpha\cos\mu t)^2 + (k\cos\alpha\cos\mu t + k\sin\alpha\cos\mu t)^2]}$$

which, when α is very small, simplifies to:

$$a(t) \cong \sqrt{[1 + 2k\sin\alpha\cos\mu t + k^2\cos^2\mu t]}$$

This can be expanded using the binomial expansion, but perhaps the exercise is getting a little tedious. One can at least see that there *will* be a fundamental while α is not zero.

FM deviation multiplication

This experiment requires a Version V.2 of the 100 kHz CHANNEL FILTERS module.

You may consider this experiment too short ? If so, then it could be extended by continuing on to the next one, entitled *FM and Bessel zeros*.

That experiment might also be considered too short, so combining them has its merits.

TRUNKS

130 kHz sinusoid from your VCO. See the answer to Q4 below.

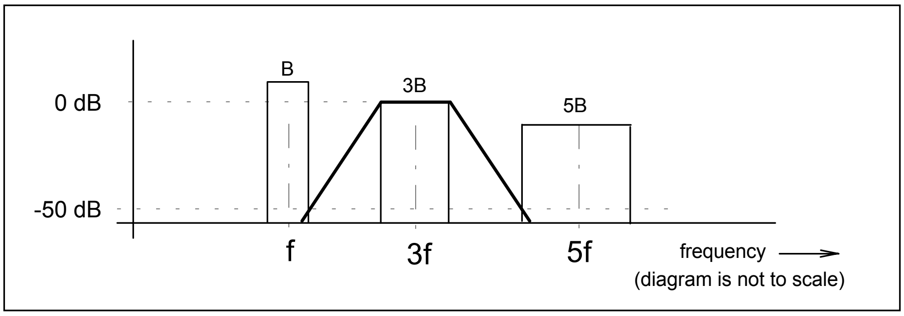
answers to tutorial questions

Q1

Q2/Q3 in practice harmonic multipliers requiring a multiplication of 9 are seldom used, since the percentage separation of the 7th and 11th (nearest unwanted odd harmonic terms) makes the filtering too difficult. For a tripler it is the fundamental and fifth which are unwanted thus making less demands upon the filter.

Working at 100 kHz and below, and with speech messages, the desired bandwidths are large fractions of the carrier frequencies, which makes the filtering quite difficult. These relationships are not common in commercial situations, where carrier frequencies are considerably higher.

The assumption is made that the bandwidth of the wanted output signal (33 kHz for the tripler) is (approximately) three times that of the input bandwidth (at 11 kHz).



harmonic multiplier output

The three rectangles represent the bands occupied by the significant components at the fundamental, third harmonic, and fifth harmonic of the 11 kHz carrier. The heavy rectangle represents the minimum requirements of the bandpass filter to select the third harmonic components from the nearest unwanted components. It is shown having an attenuation of 50 dB at the frequency where it overlaps the unwanted components. This is in excess of the 40 dB specified. This is based on the fact that, at the output of the harmonic multiplier, the amplitude of the harmonics fall in proportion to the harmonic number. Thus the components at the fundamental will be about 9 dB above those at the third harmonic (thus suggesting more than 40 dB attenuation is needed)., although those components at the upper edge of the fundamental band will be smaller than those in the centre (suggesting some relaxing of the requirements).

On the other hand a typical bandpass filter will probably find the specification on the high side the most difficult to meet, meaning that these low-side considerations may not be of importance ? What all this means is that each case must be studied on its merits.

For the case $\beta = 0.33$ then, on the 1% significant sideband criterion, from the tables 3 pairs of sideband are required to be kept. At 100 kHz ($\beta = 3$) 6 sideband pairs are significant. Notice this bandwidth increase is not in proportion to the increase of β.

For a 3 kHz top message frequency the filter bandwidths would be 2x3x3 = 18 kHz and 2x6x3 = 36 kHz. This is far wider than that available with the TIMS filters (typically 6 kHz and 20 kHz bandwidths), so the top message frequency would have to be held below about 1 kHz.

Q4 the principle which comes to mind is to use the 'BFO' principle - multiply the VCO with a fixed frequency, and take the difference frequency via a 60 kHz LPF. The VCO tunes 70 to 130 kHz. Subtracting from 100 kHz gives an output range of 0 to 30 kHz. Unsuitable.

So you - the Laboratory Manager - should supply 130 kHz at TRUNKS - from your VCO. Students can then tune their VCO through 70 to 130 kHz, and obtain 0 to 60 kHz from their 60 kHz LPF modules. Their FREQUENCY COUNTER will give them their *exact* output frequency.

FM and Bessel zeros

Refer to the previous experiment entitled *FM deviation multiplication*. It was suggested there that *this* experiment might be added on to *that* experiment.

TRUNKS

No signals required.

FM demodulation with the PLL

TRUNKS

Set up an FM signal of your choice. For qualitative measurements use a single tone as message; or perhaps a two-tone ? Otherwise speech may be preferable.

stable carrier

If you wish your generator to be based on a stable 100 kHz carrier, then Armstrong`s method is indicated. Use the generator modelled in the experiment entitled ***FM Deviation***

Multiplication. Note you will need an FM UTILITIES and a 100 kHz CHANNEL FILTERS (*type 2*) module.

modulated VCO

If you want really wideband FM you could use a VCO as the source. Although its carrier will not be stable, this may not concern you. After all, that is what the PLL is good at; ie, tracking a drifting carrier.

depth of course work

The analysis of the PLL is not a trivial matter, although the general principle of operation is simple to explain.

You can add tasks to those given, their degree of difficulty depending upon the depth of understanding you require of your students.

Likewise you can add some more demanding Tutorial Questions.

The Costas loop

TRUNKS

Two or more DSBSC, at least one of which is based on a 100 kHz carrier, and one which is not.

Costas loop filter: it would be preferable if the RC FILTER in the UTILITIES module had a longer time constant. Being near 3 Hz it is not really suitable as a loop filter in this application. A later version of this module may have a jumper on board to parallel in a larger capacitor. In the meantime make the message frequency as high as possible.

There is a preferable loop filter in the BIT CLOCK REGEN module.

The SPECTRUM UTILITIES module is very useful for measuring a DC voltage in the presence of AC components. Supply one if available.

answers to tutorial questions

Q1 the ‘squaring loop’; a PLL preceded by a squarer.

Q2 passes DC; but if too narrow acquisition takes longer.

Q3 not only acquires a carrier but adjusts it (in the I arm) to maximize the output amplitude - this is in fact the message, so Costas acts as a demodulator in addition to acquiring the carrier.

Q4 yes - the ‘third’ multiplier.

Q5 the roles of the I and Q arms will be reversed.

Q6 the roles of the I and Q arms will be reversed.

Q7 during pauses in speech there is *no signal*. It would depend upon the dynamic properties of the loop as to how long a pause could be tolerated before performance was compromised.

Q8 if the passband is not flat then sideband amplitudes will not be equal, and complete cancellation is not possible

Q9 differential encoding

Q10 would be hard to notice the difference. The loop would still endeavour to produce no output from the third multiplier. When this is achieved the wanted output would be reduced by a factor $\sin(80^0)/\sin(90^0)$.

FURTHER AND ADVANCED DIGITAL EXPERIMENTS

BER measurements in the noisy channel

The LINE-CODE ENCODER and LINE-CODE DECODER, and the DECISION MAKER, have been designed to work together as compatible interfaces between the digital input (sequence from a SEQUENCE GENERATOR) and output (sequence from the DECISION MAKER) of the system, with an analog channel in-between.

The DECISION MAKER has been optimized to operate at or around 2 kHz, and so the 2.083 kHz ‘message’ from the MASTER SIGNALS module is an ideal clock source.

A preferred method of deriving the bit clock is to use the LINE-CODE ENCODER (and its compatible partner the LINE-CODE DECODER) when working with a complete transmission system. Part of their coding job is to convert TTL level signals to TIMS-compatible analog levels. The LINE-CODE ENCODER is driven by the 8.333 kHz TTL MASTER SIGNALS clock. The LINE-CODE ENCODER provides an 8.333 kHz divided-by-four TTL output, which of course is at 2.083 kHz. This serves as the bit clock for the system, and is a rate ideal for the DECISION MAKER.

no LINE-CODE ENCODER

This first experiment with a complete transmission system does not use any line coding, but uses the LINE-CODE ENCODER and LINE-CODE DECODER in the straight-through condition, ready for line coding in a following experiment.

If you do not have LINE-CODE ENCODER and LINE-CODE DECODER modules then some modifications to the set-up for this experiment are necessary.

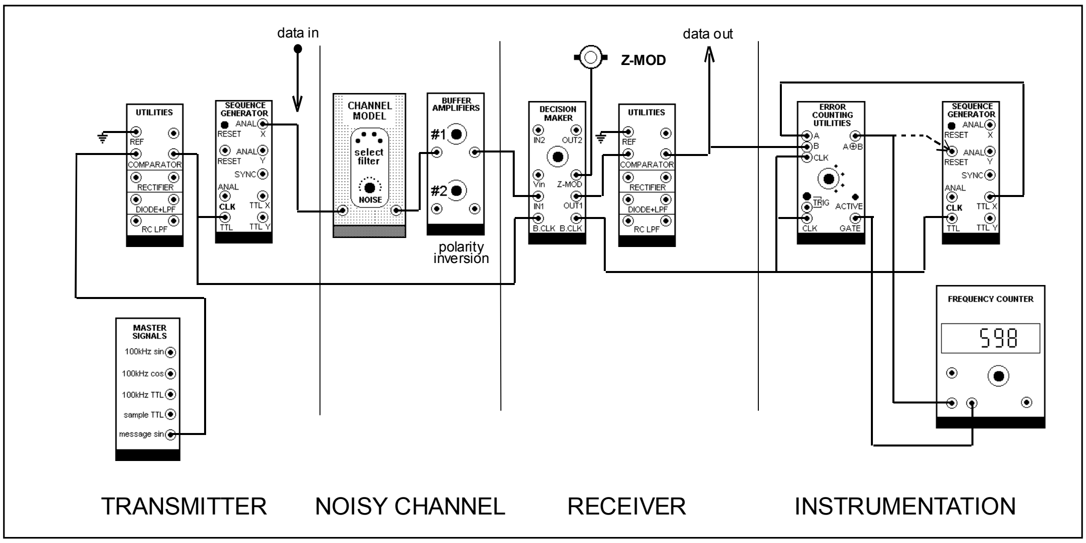
You will need two UTILITIES modules.

The SEQUENCE GENERATOR could be clocked by the 2.084 kHz sinusoidal message from MASTER SIGNALS. But a TTL version is required by the DECISION MAKER, so it is converted first to TTL by the COMPARATOR in the UTILITIES module.

The sequence input to the channel is then taken from the yellow analog output of the transmitter SEQUENCE GENERATOR. Note that this is an *inverted version* of the TTL output. This simplifies the SEQUENCE GENERATOR circuitry, but can be an occasional inconvenience, especially if not observed !

The sequence inversion must be accounted for at the X-OR gate of the ERROR COUNTING UTILITIES module. You will observe that the instrumentation SEQUENCE GENERATOR TTL stream, and that output from the DECISION MAKER, are inverted one with respect to the other. To correct for this a BUFFER is inserted at the channel output. Its gain should be set to unity.

The modified system is illustrated below.



modified system without encoder & decoder

ERROR COUNTING UTILITIES module

This module contains two sub-systems - an X-OR and a pulse counting facility. The former has been met in the experiment entitled *PRBS generation*.

These are both described in detail in the *Advanced Modules User Manual*. Condensed descriptions, suitable for this experiment, are to be found in the experiment entitled *Digital utility subsystems* under the headings *Exclusive-OR* and *Timed pulse*.

warning: make sure the default settings have been selected with the on-board switches SW1 and SW2.

TRUNKS

No signals required.

answers to tutorial questions

Q1 all SEQUENCE GENERATOR modules are identical. They are driven by the same clock. They are not influenced by the noise, or received sequence, in any way. There is no reason why they would step out of alignment.

Q2 the long sequence introduces more patterns, and exercises the system more thoroughly, thus giving a more realistic error measurement.

Q3 matching the signal to the detector threshold - which is offset from zero volts by approximately +25 mV.

BER instrumentation macro model

This Chapter is not an experiment.

Bit clock regeneration

TRUNKS

One or more line-coded signals from a SEQUENCE GENERATOR and LINE-CODE ENCODER combination should be sent via TRUNKS.

A 208 kHz sinusoid is an option (see the Experiment for details)

answers to tutorial questions

Q1 if the amplitude is varying then there must be two or more components in the signal. Assuming the largest component is at carrier frequency, smaller components can vary the amplitude and the zero crossings. Call these smaller components sidebands. These can be resolved into in-phase and quadrature pairs (or symmetrical and anti-symmetrical components). In-phase components introduce linear modulation, and do not influence the zero crossings. Quadrature pairs introduce angle modulation, non-uniform zero crossings, and timing jitter.

Q2 non-uniform zero crossings!

Q3 amplitude jitter will not influence the zero crossings

Q4

Q5

Carrier acquisition

TRUNKS

modulated

A modulated signal on a carrier of 100 kHz, and another at 50 kHz are required, from which carriers will be recovered. A carrier of 50 kHz is unusual, but convenient for the acquisition VCO, which (with the squarer) will be tuning to 100 kHz.

You should nominate which signal is which.

The signal on 100 kHz should have a carrier component present, but this is not necessary (to be avoided, in fact) for the one on 50 kHz.

To obtain 50 kHz you can divide the MASTER CLOCK by 2 and use the 60 kHz LPF. Otherwise use a VCO.

carrier

A copy of the 50 kHz carrier will be useful, but certainly not necessary, for comparison purposes. Beware, though, if the phase between recovered and stolen carriers is measured (either the 50 kHz or the 100 kHz example), that phase shifts via TRUNKS will make the answer less meaningful.

answers to tutorial questions

Q1 the SQUARER would not be required, nor the divider at the output.

Q2 there is a large DC component in the output of any squarer. This is not required. If present it could overload any following stages (here a MULTIPLIER).

Q3 not essential. But, if as might be the case in practice, the signal is accompanied by other signals, and noise, these could degrade the performance of the SQUARER in its main task.

DPSK - carrier acquisition and BER

TRUNKS

Two signals, a DPSK and a sinusoidal carrier.

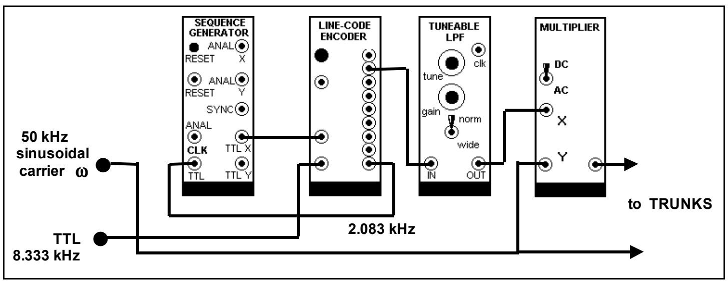
DPSK

A DPSK signal, based on a carrier of f_0 = 50 kHz.

Obtain the carrier by divide-by-2 of the 100 kHz TTL MASTER, then filtering by a 60 kHz LPF module. A TTL will not overload this module if the gain is set for an output within the TIMS ANALOG REFERENCE LEVEL limits. There will be a large DC component. This can be removed by passing the filter output via a PHASE SHIFTER module. Alternatively the following MULTIPLIER may be set to pass AC only (the NRZ-M signal should have no DC component).

The message will be from a SEQUENCE GENERATOR, set on a long sequence, and clocked at 2.083 kHz. (f_0 / 24) kHz.

A suggested model is illustrated below:



DPSK generator

carrier and bit clock

Students will not normally have a 50 kHz sinusoidal carrier signal, so this should also be sent via TRUNKS.

answers to tutorial questions

Q1 the SNR is adjusted at the DECISION MAKER input. If it is of comparable power to the signal at this point, then it will be much larger prior to bandlimiting, which occurs immediately before the DECISION MAKER. . Hence there is a danger it may overload some parts of the system between the noise source and the filter.

The noise into the MULTIPLIER is wideband, with components near 50 kHz (desired) but also around 100 kHz. These latter components will also be translated into the passband of the bandlimiting filter – this is the image response. Being white noise, it will be combined with the desired noise, and appear as – noise.

Q2 impure local oscillator means more noise (or other unwanted) components out of the MULTIPLIER.

PCM - TDM

TRUNKS

None. Speech, perhaps ? Or students may have SPEECH modules ?

WARNING

Please note that the maximum clock rate for the PCM modules is about 10 kHz. Thus it is safe to operate at the clock rate of 8.333 kHz.

PCM DECODER

See the note re this module in the PCM DEMODULATION experiment.

answers to tutorial questions

Q1 using the frame synch signal FS for oscilloscope triggering is useful for showing frames, but if the sweep circuits 'miss a beat' then adjacent frames will swap positions on the screen. If the FS signal is divided-by-2 this cannot happen. Use the divide-by-two sub-system in the BIT CLOCK REGEN module.

Block coding and decoding

TRUNKS

None

WARNING

Please note that the maximum clock rate for the block code modules is about 2 kHz, for which frequency the modules have been optimized. It is *not* safe to operate at a clock rate of 8.333 kHz.

answers to tutorial questions

Q1 assuming errors are ‘reasonable’, rather than catastrophic, then the frame must have already been identified. Thus the LSB is already without error. The error detector is looking for errors in the bits representing the message.

Q2 because of the alternating 0 and 1 pattern in the LSB position, adjacent frames will never be identical. With a DC message every *other* frame is identical. So a synch signal is needed which will look for every other frame. Dividing FS by two achieves this. If a divide-by-2 sub-system is not available, you can use a divide-by-four (see the experiment entitled *Digital utility subsystems* for other sources of digital dividers).

Block coding and coding gain

comments

Need a version 2 or above PCM DECODER module. This has an in-built message reconstruction filter.

System requires 12 slots if the optional WIDEBAND TRUE RMS METER is used.

TRUNKS

none.

Convolutional coding

The experiment is divided into two parts, A and B.

Depending on the time available each part could be considered as a separate experiment, although this is not essential.

However, it *is* essential that the experiment entitled *BER and the noisy channel* should have been completed - at another time and during a full laboratory session - before Part B be attempted. That experiment itself depends upon the successful completion of *other* experiments.

Part B uses more modules than can be accommodated in a single *TIMS 301* system. Either a second *TIMS 301*, or a *TIMS Junior*, will be required.

An alternative is to place the transmitter at TRUNKS, with stolen clocks provided, but this option has not been considered. Without a doubt it is preferable for each student to have free and easy access to both ends of the system.

Each TIMS320 DSP-DB module should already have installed in it the two EPROMs containing the convolutional decoding algorithms. The four MEMORY SELECT jumpers should be in the upper ‘A’ locations. Jumper J1 (near EPROM U5) selects the decoding algorithm for CODE 1 or CODE 2; please start in the ‘L’ position - this is CODE 1.

The alternative option of access to a PC via the front panel SERIAL LINK is not considered in this experiment. However, it is an acceptable alternative.

Although some of the information contained in the *Advanced Modules User Manual* is reproduced within the experiment text, it would be helpful if this manual was made available to students.

TRUNKS

None

TCM - trellis coding

PPM and PWM

Because almost all of the required functions are in the single INTEGRATE & DUMP module there is very little patching to be done.

There are no detailed step-by-step instructions given.

Students have to plan their own procedures. Their aim, apart from getting a feel for the signals involved, is to make sufficient measurements to enable them to answer the Tutorial Questions.

TRUNKS

None. Speech, perhaps ? Or students may have SPEECH modules ?

answers to tutorial questions

Q1 why not ?

Q2 each pulse of a monostable pulse generator can be *started* by a clock pulse (fixed position), and stopped by the PPM signal - making a variable width pulse.

Q3 more sensitive to revealing distortion.

Q4

Q5 the loop filter is too narrow for message frequencies in the range 300-3000 Hz.

QAM and 4-PSK

TRUNKS

For the students to generate and demodulate their own QPSK signal would ideally require four MULTIPLIER modules per TIMS 301 bay.

A simplification is for each group of students to generate a QPSK (two MULTIPLIER modules) and to model only one channel of the QPSK receiver (one MULTIPLIER).

If you have only two MULTIPLIER modules per student group then the solution is for you to provide a QPSK with which they can test their demodulator.

answers to tutorial questions

Q1 starting with a DSB signal, whose bandwidth is twice that of the message from which it was derived, the QAM signal adds a second channel occupying the same bandwidth and spectrum location. Twice as many message channels in the same transmission channel.

Q2 the required phase is that between the *resultant phasors* of the two DSBSC. There are typically no components at these frequencies. So a direct measurement is not possible. We make indirect measurements by noting the phase difference between the two carriers which will null each DSBSC at the output of a product demodulator.

Q3 channel discrimination is determined by the phase difference between the two (suppressed) carriers. The unwanted channel is nulled. The other channel will be of maximum amplitude (and SNR - the noise is phase insensitive) if the two DSB were at 90^0. If they were at 80^0 (say) nulling can still be achieved, but the wanted amplitude would be slightly reduced (by a factor of $\sin 80^0/\sin 90^0$) - this is quite small. Smaller deviations from 90^0 are relatively unimportant.

Q4 although only one channel demodulator was modelled, had there been two they would have acted independently. The carrier phasing adjustment of one is independent of the phasing adjustment of the other.

Q5 there being a 90^0 phase difference, the peak amplitude would be $\sqrt{2}$ volt

Q6 one less clock to recover

Q7 the modulator is based on a multiplier. Typically multipliers are susceptible to overload, which gives rise to the generation of unwanted spectral components (non-linear distortion).

Multi-level QAM and PSK

Read the *Advanced Modules User Guide,* which includes details of the following adjustments.

- set the on-board jumper J3 of the M-LEVEL ENCODER to the NORMAL position.
- set the on-board range jumper of the M-LEVEL DECODER to HI (for an 8.333 kHz clock).
- if BER measurements are to be made, check the adjustments of the on-board trimmers RV1 and RV2, according to the procedure described in the *Advanced Module User Manual*.
- make sure the Z-modulation facility has been set up to suit the oscilloscopes in use.

rack space

If each student position has a TIMS Junior as well as a TIMS 301, then there is sufficient room to model both a transmitter and a receiver. However, the experiment has been written as though such extra space is *not* available.

Even so, the complete receiving system, including noise and error rate measurement, requires 13 slots.

It is possible to manage with 12 slots, however, by noting that once the signal-to-noise ratio has beeen set at the M-LEVEL DECODER the WIDEBAND TRUE RMS METER can be removed and replaced by the ERROR COUNTING UTILITIES module, since these two are not required simultaneously.

DC offsets

You may consider it important to fine trim the DC offsets into each of the M-LEVEL DECODER modules used by the students. For details see the *Advanced Modules User Manual*.

at your discretion

The experiment brings the student to the point where the system can be set up and demonstrated. It does not describe, but leaves plenty of room for many more measurements and observations.

For example, for the various multi-level signals available:

- estimation of bandwidth
- estimation of maximum transmission speed, for a fixed bandwidth, noise free system, as judged by eye pattern quality.
- determination of maximum speed for a BER equal to or better than a specified figure for a noisy, fixed bandwidth system.

Some of these require access to both ends of the transmission system, and so more than 12 slots as provided by a single TIMS 301 rack.

Reference should be made to the use of these modulation techniques in modems, especially those used over fixed bandwidth, low-noise telephone lines.

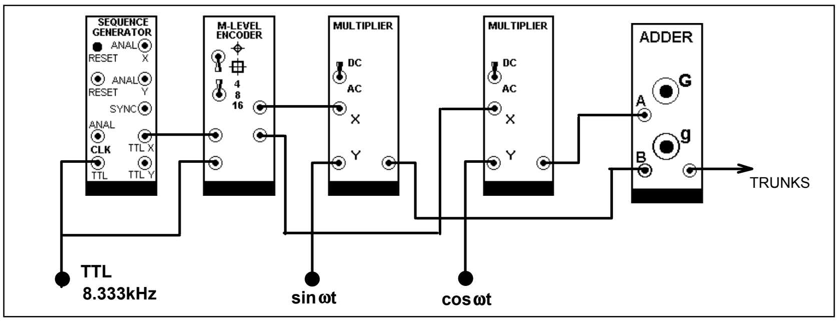
TRUNKS

At least two signals should be sent to TRUNKS. Each will require an M-LEVEL ENCODER, and two MULTIPLIER modules. Data clock of 8.333 Hz and carriers of 100 kHz.

For the first a 4-QAM signal is required (as expected by students). This gives the minimum of levels in the two paths of the demodulator output, and so makes for an easier setting of the phasing. Send two others of your choice

Each signal will use the TTL output ‘X’ from a SEQUENCE GENERATOR. Use a short sequence for the 4-QAM, longer for the others.

Use the patching shown to ensure matching relative phases at the students demodulators (as instructed in the experiment).



a 4-QAM transmitter

answers to tutorial questions

Q1 the decision point is set according to the waveform at the output of one of the filters. The same point in time is used for the decision point of the other waveform. Unless the filters are reasonably closely matched (especially with respect to delay) the decision point of the other waveform will not be optimum.

Q2 reasonable or not ? Quite OK under high signal-to-noise ratio conditions, as is demonstrated by zero errors. For marginal SNR it would best be decided by measurement (if only because not enough information is known about the response of the PHASE CHANGER). The approximate phase adjustment method could ensure that the **q** signal could be nulled from the output of the **i** channel, but not that the **i** signal could be *simultaneously* nulled from the output of the **q** channel (for example). If this is significant, it would show up in the BER under poor SNR conditions.

Q3 if wideband noise enters the input to the receiver not only will components at around 100 kHz be frequency translated to baseband, along with the incoming signal, *but also* noise components around 200 kHz. This would degrade the performance of the receiver from that achievable with an input, *image rejection*, filter. But in the measurement situation we have control over the noise - we merely have to add about half as much at the input to the receiver as would otherwise be required for a given amount at the detector input (where SNR is to be measured). This is a problem, but can be accounted for.

Note that a 2 dB change at the NOISE GENERATOR attenuator would make more than a 2 dB change at the measurement point. How much more ?

In an attempt to obtain enough noise at the detector input it is tempting to amplify it at source. But in the present example this would be *before* bandlimiting. Beware of overloading the multipliers.

Spread spectrum - DSSS & CDMA

TRUNKS

See experiment.

A DSSS signal, where the PN sequence is derived from a VCO near 100 kHz. This will be a signal sharing the channel, but not part of a CDMA system.

answers to tutorial questions

Q1

Q2 each of the spectral lines of the PN sequence locates the carrier of a DSBSC signal. Each of the sidebands of this DSBSC is a frequency translated version of the baseband message (one erect, the other inverted). Depending upon the spacing of the lines in the PN sequence these DSBSC may or may not overlap. If they do *not* overlap the job of an interceptor (eavesdropper) is simplified.

Q3 there are two DSBSC in a phase division multiplexed signal. The two DSBSC are phased nominally at 90^0 at the transmitter. The demodulator recovers both channels simultaneously, so that their messages appear *added* at the demodulator output. Their relative amplitudes are determined by the phase of the demodulating carrier. When the amplitude of one message is reduced to zero (nulled) that of the other is maximized. In this way one or the other channel can be recovered independent of the other.

Thus it is not the wanted channel which is maximized, but the unwanted channel which is minimized.

There cannot be more than three or more channels in the PDM system, since it would be necessary then to minimize two or more simultaneously, which would be impossible.

In the spread spectrum signal there are literally thousands of DSBSC signals. The correct demodulating sequence contains thousands of individual carriers, all at the correct frequency. The identical messages from each of these DSBSC appear added at the demodulator output. If the relative phasing of each of the components in the demodulating sequence is correct, these contributions are combined so as to optimize the resultant amplitude. Otherwise they will combine to produce a much smaller resultant.

The selection process is an *enhancement* of the wanted message, unlike the PDM case, where it is a *minimization* of the unwanted message.

Q4

Q5

Q6 more confusing to the interceptor. See answer to Q2

Q7

Q8 one less clock to recover; influences the spectrum

Digital utility sub-systems

This Chapter is not a conventional experiment.