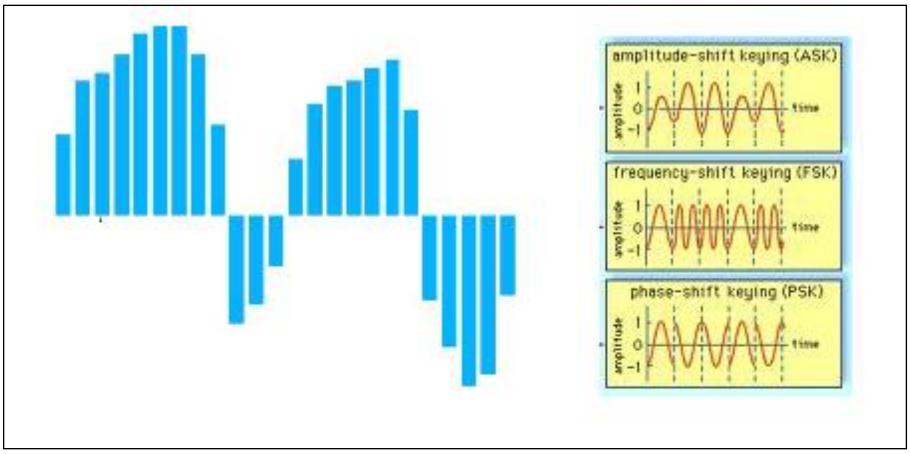
Department of Electrical and Telecommunications Engineering Technology



Laboratory Manual

EET 3202: Principles of Communications Systems

Prepared by

Prof. Kouar and Prof. Hossain

(Fall 2013)

Table of Contents

Week	Lab	Page
1	1. Modelling an equation	18
2 & 3	2. DSB-SC generation 3. Amplitude Modulation 4. Envelope recovery	30 43 54
4	5. Introduction to FM using a VCO	65
5	6. FM demodulation with the PLL	74
6 & 7	7. Sampling with Sample and Hold	79
8	8. PCM coding	85
9	9. PCM decoding	95
10	10. Delta modulation	104
11	11. Delta demodulation	115
12	12. Line coding	120
13	13. ASK-Amplitude Shift Keying	127
14	14. Block coding and decoding	136
15	Final exam	

Acknowledgement:

We like to thank Emona Technologies (http://www.qpsk.com/) for providing us with the lab manuals.

INTRODUCTION TO MODELLING WITH TIMS

model building

With TIMS you will be building models. These models will most often be hardware realizations of the block diagrams you see in a text book, or have designed yourself. They will also be representations of equations, which themselves can be depicted in block diagram form.

Whatever the origin of the model, it can be patched up in a very short time. The next step is to adjust the model to perform as expected. It is perfectly true that you might, on occasions, be experimenting, or just ‘doodling’, not knowing what to expect. But in most cases your goal will be quite clear, and this is where a systematic approach is recommended.

If you follow the steps detailed in the first few experiments you will find that the models are adjusted in a systematic manner, so that each desired result is obtained via a complete understanding of the purpose and aim of the intermediate steps leading up to it.

why have patching diagrams ?

Many of the analog experiments, and all of the digital experiments, display patching diagrams. These give all details of the interconnections between modules, to implement a model of the system under investigation.

> *It is **not expected** that a glance at the patching diagram will reveal the nature of the system being modelled.*

The patching diagram is presented as firm evidence that a model of the system can be created with TIMS.

> *The functional purpose of the system is revealed through the block diagram which precedes the patching diagram.*

It is the block diagram which you should study to gain insight into the workings of the system.

If you fully understand the block diagram you should not need the patching diagram, except perhaps to confirm which modules are required for particular operations, and particular details of functionality. These is available in the ***TIMS User Manual***.

You may need an occasional glance at the patching diagram for confirmation of a particular point.

Try to avoid patching up 'mechanically', according to the patching diagram, without thought to what you are trying to achieve.

organization of experiments

Each of the experiments in this Text is divided into three parts.

1. The first part is generally titled *PREPARATION.* This part should be studied *before* the accompanying laboratory session.
2. The second part describes the experiment proper. Its title will vary. You will find the experiment a much more satisfying experience if you arrive at the laboratory well prepared, rather than having to waste time finding out what has to be done at the last moment. Thus read this part *before* the laboratory session.
3. The third part consists of TUTORIAL QUESTIONS. Generally these questions will be answered after the experimental work is completed, but it is a good idea to read them *before* the laboratory session, in case there are special measurements to be made.

While performing an experiment you should always have access to the ***TIMS*** user manuals - namely the ***TIMS User Manual*** (fawn cover) which contains information about the modules in the TIMS Basic Set of modules, and the ***TIMS Advanced Modules and TIMS Special Applications Modules User Manual*** (red cover).

who is running this experiment ?

These experiments and their Tasks are merely suggestions as to how you might go about carrying out certain investigations. In the final assessment it is *you* who are running the experiment, and *you* must make up your mind as to how you are going to do it. You can do this best if you read about it beforehand.

If you do not understand a particular instruction, consider what it is you have been trying to achieve up to that point, and then do it your way.

early experiments

The first experiment assumes no prior knowledge of telecommunications - it is designed to introduce you to TIMS, and to illustrate the previous remarks about being systematic. The techniques learned will be applied over and over again in later work.

The next few experiments are concerned with analog modulation and demodulation.

modulation

One of the many purposes of *modulation* is to convert a message into a form more suitable for transmission over a particular medium.

The analog modulation methods to be studied will generally transform the analog message signal in the audio spectrum to a higher location in the frequency spectrum.

The digital modulation methods to be studied will generally transform a binary data stream (the message), at baseband [1] frequencies, to a different format, and then may or may not translate the new form to a higher location in the frequency spectrum.

It is much easier to radiate a high frequency (HF) signal than it is a relatively low frequency (LF) audio signal. In the TIMS environment the particular part of the spectrum chosen for HF signals is centred at 100 kHz.

It is necessary, of course, that the reverse process, *demodulation*, can be carried out - namely, that the message may be recovered from the modulated signal upon receipt following transmission.

messages

Many models will be concerned with the transmission or reception of a message, or a signal carrying a message. So TIMS needs suitable messages. These will vary, depending on the system.

analog messages

The transmission of speech is often the objective in an analog system.

High-fidelity speech covers a wide frequency range, say 50 Hz to 15 kHz, but for communications purposes it is sufficient to use only those components which lie in the audio frequency range 300 to 3000 Hz - this is called ‘band limited speech’. Note that frequency components have been removed from both the low and the high frequency end of the message spectrum. This is bandpass filtering. Intelligibility suffers if only the high frequencies are removed.

Speech is not a convenient message signal with which to make simple and precise measurements. So, initially, a single tone (sine wave) is used. This signal is more easily accommodated by both the analytical tools and the instrumentation and measuring facilities.

The frequency of this tone can be chosen to lie within the range expected in the speech, and its peak amplitude to match that of the speech. The simple tone can then be replaced by a two-tone test signal, in which case intermodulation tests can be carried out [2].

When each modulation or demodulation system has been set up quantitatively using a single tone as a message (or, preferably with a two-tone test signal), a final qualitative check can be made by replacing the tone with a speech signal. The peak amplitude of the speech should be adjusted to match that of the tone.

[1] defined later

[2] the two-tone test signal is introduced in the experiment entitled ‘***Amplifier overload***’.

Both listening tests (in the case of demodulation) and visual examination of the waveforms can be very informative.

digital messages

The transmission of binary sequences is often the objective of a digital communication system. Of considerable interest is the degree of success with which this transmission is achieved. An almost universal method of describing the quality of transmission is by quoting an error rate [3].

If the sequence is one which can take one of two levels, say 0 and 1, then an error is recorded if a 0 is received when a 1 was sent, or a 1 received when a 0 was sent. The bit error rate is measured as the number of errors as a proportion of total bits sent.

To be able to make such a measurement it is necessary to know the exact nature of the original message. For this purpose a known sequence needs to be transmitted, a copy of which can be made available at the receiver for comparison purposes. The known sequence needs to have known, and useful, statistical properties - for example, a ‘random’ sequence. Rather simple generators can be implemented using shift registers, and these provide sequences of adjustable lengths. They are known as pseudo-random binary sequence (PRBS) generators. TIMS provides you with just such a SEQUENCE GENERATOR module. You should refer to a suitable text book for more information on these.

bandwidths and spectra

Most of the signals you will be examining in the experiments to follow have well defined bandwidths. That is, in most cases it is possible to state quite clearly that all of the energy of a signal lies between frequencies f_1 and f_2 Hz, where $f_1 < f_2$.

- the *absolute bandwidth* of such a signal is defined as $(f_2 - f_1)$ Hz.

It is useful to define the number of octaves a signal occupies. The octave measure for the above signal is defined as

$$\text{octaves} = \log_2(f_2 / f_1)$$

Note that the octave measure is a function of the *ratio* of two frequencies; it says nothing about their *absolute* values.

- a *wideband signal* is generally considered to be one which occupies one or more octaves.
- a *narrowband signal* is one which occupies a small fraction of an octave. Another name, used interchangeably, is a *bandpass* signal.

An important observation can be made about a narrowband signal; that is, it can contain no harmonics.

- a *baseband signal* is one which extends from DC (so $f_1 = 0$) to a finite frequency f_2. It is thus a *wideband* signal.

Speech, for communications, is generally bandlimited to the range 300 to 3000 Hz. It thus has a bandwidth in excess of 3 octaves. This is considered to be a wideband signal. After modulation, to a higher part of the spectrum, it becomes a narrowband signal, but note that *its absolute bandwidth remains unchanged*.

[3] the corresponding measurement in an analog system would be the signal-to-noise ratio (relatively easy to measure with instruments), or, if speech is the message, the ‘intelligibility’; not so easy to define, let alone to measure.

This reduction from a wideband to a narrowband signal is a linear process; it can be reversed. In the context of communications engineering it involves *modulation*, or *frequency translation.*

You will meet all of these signals and phenomena when working with TIMS.

measurement

The bandwidth of a signal can be measured with a SPECTRUM ANALYSER. Commercially available instruments typically cover a wide frequency range, are very accurate, and can perform a large number of complex measurements. They are correspondingly expensive.

TIMS has no spectrum analyser as such, but can model one (with the TIMS320 DSP module), or in the form of a simple WAVE ANALYSER with TIMS analog modules. See the experiment entitled ***Spectrum analysis - the WAVE ANALYSER*** (within *Volume A2 - Further & Advanced Analog Experiments*).

Without a spectrum analyser it is still possible to draw conclusions about the location of a spectrum, by noticing the results when attempting to pass it through filters of different bandwidths. There are several filters in the TIMS range of modules. See Appendix A, and also the ***TIMS User Manual***.

graphical conventions

representation of spectra

It is convenient to have a graphical method of depicting spectra. In this work we do not get involved with the Fourier transform, with its positive and negative frequencies and double sided spectra. Elementary trigonometrical methods are used for analysis. Such methods are more than adequate for our purposes.

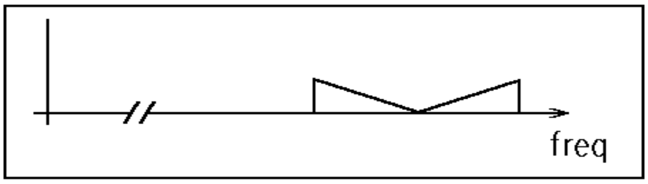
When dealing with speech the mathematical analysis is dropped, and descriptive methods used. These are supported by graphical representations of the signals and their spectra.

In the context of modulation we are constantly dealing with *sidebands*, generally derived from a baseband message of finite bandwidth. Such finite bandwidth signals will be represented by triangles on the spectral diagrams.

The steepness of the slope of the triangle has no special significance, although when two or more sidebands, from different messages, need to be distinguished, each can be given a different slope.

a baseband signal (eg., a message)

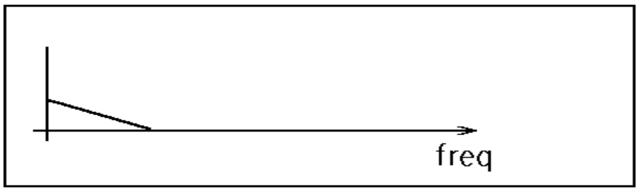
Although speech does not have a DC component, the triangle generally extends down to zero (the origin) of the frequency scale (rather than being truncated just before it). For the special case in which a baseband signal does have a DC component the triangle convention is sometimes modified slightly by adding a vertical line at the zero-frequency end of the triangle.



a DSBSC

The *direction* of the slope is important. Its significance becomes obvious when we wish to draw a modulated signal. The figure above shows a double sideband suppressed carrier (DSBSC) signal.

Note that there are TWO triangles, representing the individual lower and upper sidebands. They slope towards the same point; this point indicates the location of the (suppressed) carrier frequency.

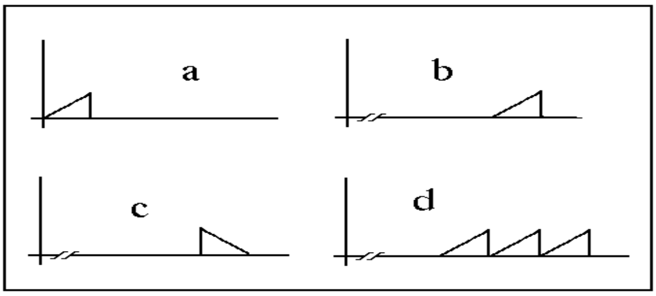


an inverted baseband signal

The *orientation* is important. If the same message was so modulated that it could be represented in the frequency spectrum as in the figure above, then this means:

- the signal is located in the baseband part of the spectrum
- spectral components have been transposed, or inverted; frequency components which were originally above others are now below them.
- since the signal is at baseband it would be audible (if converted with an electric to acoustic transducer - a pair of headphones, for example), but would be unintelligible. You will be able to listen to this and other such signals in TIMS experiments to come.

It is common practice to use the terms *erect* and *inverted* to describe these bands.



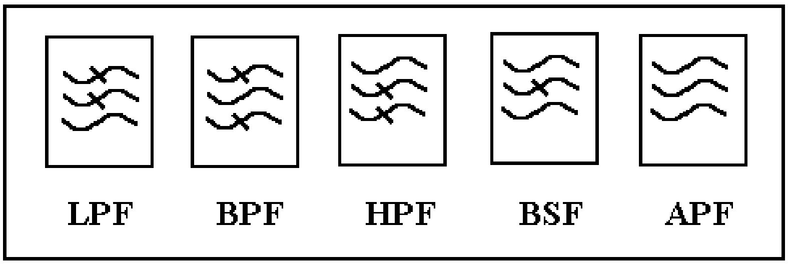
In the Figure above, a message (a) is frequency translated to become an upper single sideband (b), and a lower single sideband (c). A three-channel frequency division multiplexed (FDM) signal is also illustrated (d).

Note that these spectral diagrams do not show any phase information.

Despite all the above, be prepared to accept that these diagrams are used for purposes of *illustration*, and different authors use their own variations. For example, some slope their triangles in the opposite sense to that suggested here.

filters

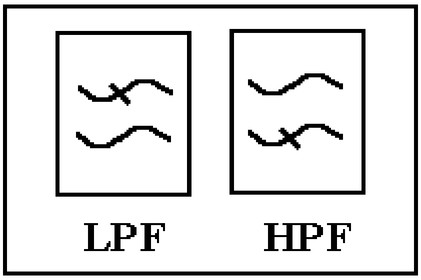
In a block diagram, there is a simple technique for representing filters. The frequency spectrum is divided into three bands - low, middle, and high - each represented by part of a sinewave. If a particular band is blocked, then this is indicated by an oblique stroke through it. The standard responses are represented as in the Figure below.



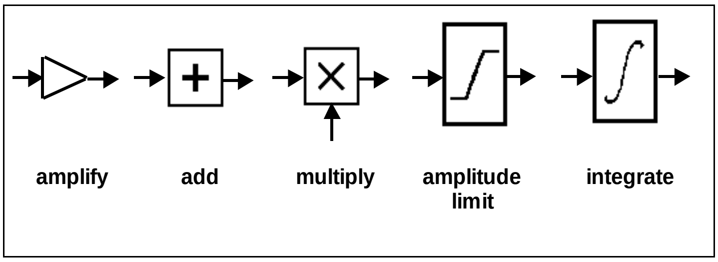
block-diagrammatic representations of filter responses

The filters are, respectively, lowpass, bandpass, highpass, bandstop, and allpass.

In the case of lowpass and highpass responses the diagrams are often further simplified by the removal of one of the cancelled sinewaves, the result being as in the figure opposite.



other functions



some analog functions

measuring instruments

the oscilloscope - time domain

The most frequently used measuring facility with TIMS is the oscilloscope. In fact the vast majority of experiments can be satisfactorily completed with no other instrument.

Any general purpose oscilloscope is ideal for all TIMS experiments. It is intended for the display of signals in the time domain [4]. It shows their waveforms - their shapes, and amplitudes

From the display can be obtained information regarding:

- waveform shape
- waveform frequency - by calculation, using time base information
- waveform amplitude - directly from the display
- system linearity - by observing waveform distortion
- an *estimate* of the bandwidth of a complex signal; eg, from the sharpness of the corners of a square wave

When concerned with amplitude information it is customary to record either:

- the *peak-to-peak* amplitude
- the *peak* amplitude

of the waveform visible on the screen.

Unless the waveform is a simple sinewave it is always important to record the *shape* of the waveform also; this can be:

1. as a sketch (with time scale), and annotation to show clearly what amplitude has been measured.
2. as an analytic expression, in which case the parameter recorded must be clearly specified.

the rms voltmeter

The TIMS WIDEBAND TRUE RMS METER module is essential for measurements concerning *power*, except perhaps for the simple case when the signal is one or two sinewaves. It is particularly important when the measurement involves *noise*.

Its bandwidth is adequate for all of the signals you will meet in the TIMS environment.

An experiment which introduces the WIDEBAND TRUE RMS METER, is entitled *Power measurements*. Although it appears at the end of this Volume, it could well be attempted at almost any time.

the spectrum analyser - frequency domain

The identification of the spectral composition of a signal - its components in the frequency domain - plays an important part when learning about communications.

[4] but with adaptive circuitry it can be modified to display frequency-domain information

Unfortunately, instruments for *displaying* spectra tend to be far more expensive than the general purpose oscilloscope.

It is possible to identify and measure the individual spectral components of a signal using TIMS modules.

Instruments which identify the spectral components on a component-by-component basis are generally called *wave analysers*. A model of such an instrument is examined in the experiment entitled ***Spectrum analysis - the WAVE ANALYSER*** in *Volume A2 - Further & Advanced Analog Experiments*.

Instruments which identify the spectral components of a signal and *display* the spectrum are generally called *spectrum analysers*. These instruments tend to be more expensive than wave analysers. Something more sophisticated is required for their modelling, but this is still possible with TIMS, using the digital signals processing (DSP) facilities - the TIMS320 module can be programmed to provide spectrum analysis facilities.

Alternatively the distributors of TIMS can recommend other affordable methods, compatible with the TIMS environment.

oscilloscope - triggering

synchronization

As is usually the case, to achieve 'text book like' displays, it is important to choose an appropriate signal for oscilloscope triggering. This trigger signal is almost never the signal being observed ! The recognition of this point is an important step in achieving stable displays.

This chosen triggering signal should be connected *directly* to the oscilloscope sweep synchronizing circuitry. Access to this circuitry of the oscilloscope is available via an input socket other than the vertical deflection amplifier input(s). It is typically labelled '***ext. trig***' (external trigger), '***ext. synch***' (external synchronization), or similar.

sub-multiple frequencies

If two or more periodic waveforms are involved, they will only remain stationary with respect to each other if the frequency of one is a sub-multiple of the other. This is seldom the case in practice, but can be made so in the laboratory. Thus TIMS provides, at the MASTER SIGNALS module, a signal of 2.083 kHz (which is 1/48 of the 100 kHz system clock), and another at 8.333 kHz (1/12 of the system clock).

which channel ?

Much time can be saved if a consistent use of the SCOPE SELECTOR is made. This enables quick changes from one display to another with the flip of a switch. In addition, *channel identification* is simplified if the habit is adopted of consistently locating the trace for CH1 *above* the trace for CH2.

Colour coded patching leads can also speed trace identification.

what you see, and what you don`t

Instructions such as '*adjust the phase until there is no output*', or '*remove the unwanted signal with a suitable filter*' will be met from time to time.

These instructions seldom result in the amplitude of the signal in question being reduced to zero. Instead, what is generally meant is '*reduce the amplitude of the signal until it is no longer of any significance*'.

Significance here is a relative term, made with respect to the system signal-to-noise ratio (SNR). All systems have a background noise level (noise *threshold*, noise *floor*), and signals (wanted) within these systems must over-ride this noise (unwanted).

TIMS is designed to have a 'working level', the TIMS ANALOG REFERENCE LEVEL, of about 4 volts peak-to-peak. The system noise level is claimed to be at least 100 times below this [5].

When using an oscilloscope as a measuring instrument with TIMS, the vertical sensitivity is typically set to about 1 volt/cm. Signals at the reference level fit nicely on the screen. If they are too small it is wise to increase them if possible (and appropriate), to over-ride the system noise; or if larger to reduce them, to avoid system overload.

When they are attenuated by a factor of 100, and if the oscilloscope sensitivity is not changed, they *appear* to be '*reduced to zero*'; and in relative terms this is so.

If the sensitivity of the oscilloscope is increased by 100, however, the screen will no longer be empty. There will be the system noise, and perhaps the signal of interest is still visible. Engineering judgement must then be exercised to evaluate the significance of the signals remaining.

overload

If wanted signal levels within a system fall 'too low' in amplitude, then the signal-to-noise ratio (SNR) will suffer, since internal circuit noise is independent of signal level.

If signal levels within a system rise 'too high', then the SNR will suffer, since the circuitry will overload, and generate extra, unwanted, distortion components; these distortion components are signal level dependent. In this case the noise is derived from distortion of the signal, and the degree of distortion is usually quoted as signal-to-distortion ratio (SDR).

Thus analog circuit design includes the need to maintain signal levels at a pre-defined working level, being 'not to high' and 'not too low', to avoid these two extremes.

These factors are examined in the experiment entitled ***Amplifier overload*** within *Volume A2 - Further & Advanced Analog Experiments*.

The TIMS working signal level, or TIMS ANALOG REFERENCE LEVEL, has been set at 4 volts peak-to-peak. Modules will generally run into non-linear operation when this level is exceeded by say a factor of two. The background noise of the TIMS system is held below about 10 mV - this is a fairly loose statement, since this level is dependent upon the bandwidth over which the noise is measured, and the model being examined at the time. A general statement would be to say that TIMS endeavours to maintain a SNR of better than 40 dB for all models.

overload of a narrowband system

Suppose a channel is narrowband. This means it is deliberately bandlimited so that it passes signals in a narrow (typically much less than an octave [6]) frequency range only. There are many such circuits in a communications system.

---

[5] TIMS claims a system signal-to-noise ratio of better than 40 dB

If this system overloads on a single tone input, there will be unwanted harmonics generated. But these will not pass to the output, and so the overload may go unnoticed. With a more complex input - say two or more tones, or a speech-related signal - there will be, in addition, unwanted intermodulation components generated. Many of these *will* pass via the system, thus revealing the existence of overload. In fact, the two-tone test signal should always be used in a narrowband system to investigate overload.

the two-tone test signal

A two-tone test signal consists of two sine waves added together ! As discussed in the previous section, it is a very useful signal for testing systems, especially those which are of narrow-bandwidth. The properties of the signal depend upon:

- the frequency ratio of the two tones.
- the amplitude ratio of the two tones.

For testing narrowband communication systems the two tones are typically of near-equal frequency, and of identical amplitude. A special property of this form of the signal is that its shape, as seen in the time domain, is very well defined and easily recognisable [7].

After having completed the early experiments you will recognise this shape as that of the double sideband suppressed carrier (DSBSC) signal.

If the system through which this signal is transmitted has a non-linear transmission characteristic, then this will generate extra components. The presence of even small amounts of these components is revealed by a change of shape of the test signal.

Fourier series and bandwidth estimation

Fourier series analysis of periodic signals reveals that:

- it is possible, by studying the symmetry of a signal, to predict the presence or absence of a DC component.
- if a signal is other than sinusoidal, it will contain *more than* one harmonic component of significance.
- if a signal has sharp discontinuities, it is likely to contain *many* harmonic components of significance
- some special symmetries result in all (or nearly all) of the ODD (or EVEN) harmonics being absent.

With these observations, and more, it is generally easy to make an engineering estimate of the bandwidth of a periodic signal.

---

[6] defined above

[7] the assumption being that the oscilloscope is set to sweep across the screen over a few periods of the *difference* frequency.

multipliers and modulators

The modulation process requires *multiplication*. But a pure MULTIPLIER is seldom found in communications equipment. Instead, a device called a MODULATOR is used.

In the TIMS system we generally use a MULTIPLIER, rather than a MODULATOR, when multiplication is called for, so as not to become diverted by the side effects and restrictions imposed by the latter.

In commercial practice, however, the purpose-designed MODULATOR is generally far superior to the unnecessarily versatile MULTIPLIER.

multipliers

An ideal multiplier performs as a multiplier should ! That is, if the two time-domain functions x(t) and y(t) are multiplied together, then we expect the result to be x(t)·y(t), no more and no less, and no matter what the nature of these two functions. These devices are called *four quadrant* multipliers.

There are practical multipliers which approach this ideal, with one or two engineering qualifications. Firstly, there is always a restriction on the bandwidth of the signals x(t) and y(t).

There will inevitably be extra (unwanted) terms in the output (noise, and particularly distortion products) due to practical imperfections.

Provided these unwanted terms can be considered 'insignificant', with respect to the magnitude of the wanted terms, then the multiplier is said to be 'ideal'. In the TIMS environment this means they are at least 40 dB below the TIMS ANALOG REFERENCE LEVEL [8].

Such a multiplier is even said to be linear. That is, from an engineering point of view, it is performing as expected.

In the mathematical sense it is not linear, since the mathematical definition of a linear circuit includes the requirement that no new frequency components are generated when it performs its normal function. But, as will be seen, multiplication always generates new frequency components.

DC off-sets

One of the problems associated with analog circuit design is minimization of unwanted DC off-sets. If the signals to be processed have no DC component (such as in an audio system) then stages can be AC coupled, and the problem is overcome. In the TIMS environment module bandwidths must extend to DC, to cope with all possible conditions; although more often than not signals have no intentional DC component.

In a complex model DC offsets can accumulate - but in most cases they can be recognised as such, and accounted for appropriately. There is one situation, however, where they can cause much more serious problems by generating *new components* - and that is when *multiplication* is involved.

With a MULTIPLIER the presence of an unintentional DC component at one input will produce new components at the output. Specifically, each component at the other input will be multiplied by this DC component - a constant - and so a scaled version will appear at the output [9].

[8] defined under 'what you see and what you don`t'

[9] this is the basis of a voltage controlled amplifier - VCA

To overcome this problem there is an option for AC coupling in the MULTIPLIER module. It is suggested that the DC mode be chosen only when the signals to be processed actually have DC components; otherwise use AC coupling.

modulators

In communications practice the circuitry used for the purpose of performing the multiplying function is not always ideal in the *four quadrant multiplier* sense; such circuits are generally called *modulators*.

Modulators generate the *wanted* sum *or* difference products but in many cases the input signals will also be found in the output, along with other *unwanted* components at significant levels. Filters are used to remove these unwanted components from the output (alternatively there are 'balanced' modulators. These have managed to eliminate either one or both of the original signals from the output).

These modulators are restricted in other senses as well. It is allowed that one of the inputs can be complex (ie., two or more components) but the other can only be a single frequency component (or appear so to be - as in the switching modulator). This restriction is of no disadvantage, since the vast majority of modulators are required to multiply a complex signal by a single-component carrier.

Accepting restrictions in some areas generally results in superior performance in others, so that in practice it is the *switching modulator*, rather than the idealized *four quadrant multiplier*, which finds universal use in communications electronics.

Despite the above, TIMS uses the four quadrant multiplier in most applications where a modulator might be used in practice. This is made possible by the relatively low frequency of operation, and modest linearity requirements

envelopes

Every narrowband signal has an envelope, and you probably have an idea of what this means.

Envelopes will be examined first in the experiment entitled ***DSB generation*** in this Volume.

They will be defined and further investigated in the experiments entitled ***Envelopes*** within this *Volume*, and ***Envelope recovery*** within *Volume A2 - Further & Advanced Analog Experiments*.

extremes

Except for a possible frequency scaling effect, most experiments with TIMS will involve realistic models of the systems they are emulating. Thus message frequencies will be 'low', and carrier frequencies 'high'. But these conditions need not be maintained. TIMS is a very flexible environment.

It is always a rewarding intellectual exercise to imagine what would happen if one or more of the 'normal' conditions was changed severely [10].

It is then even more rewarding to confirm our imaginings by actually modelling these unusual conditions. TIMS is sufficiently flexible to enable this to be done in most cases.

For example: it is frequently stated, for such-and-such a requirement to be satisfied, that it is necessary that '$x_1 >> x_2$'. Quite often x_1 and x_2 are frequencies - say a carrier and a message frequency; or they could be amplitudes.

You are strongly encouraged to expand your horizons by questioning the reasons for specifying the conditions, or restrictions, within a model, and to consider, and then examine, the possibilities when they are ignored.

analog or digital ?

What is the difference between a digital signal and an analog signal ? Sometimes this is not clear or obvious.

In TIMS digital signals are generally thought of as those being compatible with the TTL standards. Thus their amplitudes lie in the range 0 to +5 volts. They come from, and are processed by, modules having RED output and input sockets.

It is sometimes necessary, however, to use an analog filter to bandlimit these signals. But their large DC offsets would overload most analog modules, . Some digital modules (eg, the SEQUENCE GENERATOR) have anticipated this, and provide an analog as well as a digital (TTL) output. This analog output comes from a YELLOW socket, and is a TTL signal with the DC component removed (ie, DC shifted).

SIN or COS ?

Single frequency signals are generally referred to as sinusoids, yet when manipulating them trigonometrically are often written as cosines. How do we obtain $\cos\omega t$ from a sinusoidal oscillator !

There is no difference in the *shape* of a sinusoid and a cosinusoid, as observed with an oscilloscope. A sinusoidal oscillator can just as easily be used to provide a cosinusoid. What we call the signal (sin or cos) will depend upon the time reference chosen.

Remember that $\cos\omega t = \sin(\omega t + \pi/2)$

Often the time reference is of little significance, and so we choose sin or cos, in any analysis, as is convenient.

[10] for an entertaining and enlightening look at the effects of major changes to one or more of the physical constants, see G. Gamow; *Mr Tompkins in Wonderland* published in 1940, or easier *Mr. Tompkins in Paperback*, Cambridge University Press, 1965.

the ADDER - G and g

Refer to the ***TIMS User Manual*** for a description of the ADDER module. Notice it has two input sockets, labelled 'A' and 'B'.

In many experiments an ADDER is used to make a linear sum of two signals a(t) and b(t), of amplitudes **A** and **B** respectively, connected to the inputs A and B respectively. The proportions of these signals which appear at the ADDER output are controlled by the front panel gain controls **G** and **g**.

The amplitudes **A** and **B** of the two input signals are seldom measured, nor the magnitudes **G** and **g** of the adjustable gains.

Instead it is the magnitudes **GA** and **gB** which are of more interest, and these are measured directly at the ADDER output. The measurement of **GA** is made when the patch lead for input **B** is removed; and that of **gB** is measured when the patch lead for input **A** is removed.

When referring to the two inputs in this text it would be formally correct to name them as 'the input A' and 'the input B'. This is seldom done. Instead, they are generally referred to as 'the input **G**' and 'the input **g**' respectively (or sometimes just **G** and **g)**. This should never cause any misunderstanding. If it does, then it is up to you, as the experimenter, to make an intelligent interpretation.

abbreviations

This list is not exhaustive. It includes only those abbreviations used in this Text.

abbreviation	***meaning***
AM	amplitude modulation
ASK	amplitude shift keying (also called OOK)
BPSK	binary phase shift keying
CDMA	code division multiple access
CRO	cathode ray oscilloscope
dB	decibel
DPCM	differential pulse code modulation
DPSK	differential phase shift keying
DSB	double sideband (in this text synonymous with DSBSC)
DSBSC	double sideband suppressed carrier
DSSS	direct sequence spread spectrum
DUT	device under test
ext. synch.	external synchronization (of oscilloscope). ‘ext. trig.’ preferred
ext. trig.	external trigger (of an oscilloscope)
FM	frequency modulation
FSK	frequency shift keying
FSD	full scale deflection (of a meter, for example)
IP	intermodulation product
ISB	independent sideband
ISI	intersymbol interference
LSB	*analog*: lower sideband *digital*: least significant bit
MSB	most significant bit
NBFM	narrow band frequency modulation
OOK	on-off keying (also called ASK)
PAM	pulse amplitude modulation
PCM	pulse code modulation
PDM	pulse duration modulation (see PWM)
PM	phase modulation
PPM	pulse position modulation
PRK	phase reversal keying (also called PSK)
PSK	phase shift keying (also called PRK - see BPSK)
PWM	pulse width modulation (see PDM)
SDR	signal-to-distortion ratio
SNR	signal-to-noise ratio
SSB	single sideband (in this text is synonymous with SSBSC)
SSBSC	single sideband suppressed carrier
SSR	sideband suppression ratio
TDM	time division multiplex
THD	total harmonic distortion
VCA	voltage controlled amplifier
WBFM	wide band frequency modulation

list of symbols

The following symbols are used throughout the text, and have the following meanings

$a(t)$ a time varying amplitude

α, ϕ, φ, phase angles

β deviation, in context of PM and FM

δf a small frequency increment

$\Delta\phi$ peak phase deviation

δt a small time interval

$\phi(t)$ a time varying phase

m in the context of envelope modulation, the *depth* of modulation

μ a low frequency (rad/s); typically that of a message ($\mu << \omega$).

ω a high frequency (rad/s); typically that of a carrier ($\omega >> \mu$)

$y(t)$ a time varying function

1. MODELLING AN EQUATION

ACHIEVEMENTS: *a familiarity with the TIMS modelling philosophy; development of modelling and experimental skills for use in future experiments. Introduction to the ADDER, AUDIO OSCILLATOR, and PHASE SHIFTER modules; also use of the SCOPE SELECTOR and FREQUENCY COUNTER.*

PREREQUISITES: *a desire to enhance one's knowledge of, and insights into, the phenomena of telecommunications theory and practice.*

preparation

This experiment assumes no prior knowledge of telecommunications. It illustrates how TIMS is used to model a mathematical equation. You will learn some experimental techniques. It will serve to introduce you to the TIMS system, and prepare you for the more serious experiments to follow.

In this experiment you will model a simple trigonometrical equation. That is, you will demonstrate in hardware something with which you are already familiar analytically.

an equation to model

You will see that what you are to do experimentally is to demonstrate that two AC signals of the *same* frequency, *equal* amplitude and *opposite* phase, when added, will sum to zero.

This process is used frequently in communication electronics as a means of removing, or at least minimizing, unwanted components in a system. You will meet it in later experiments.

The equation which you are going to model is:

$$y(t) = V_1 \sin(2\pi f_1 t) + V_2 \sin(2\pi f_2 t + \alpha) \qquad \text{........ 1}$$

$$= v_1(t) + v_2(t) \qquad \text{........ 2}$$

Here y(t) is described as the sum of two sine waves. Every young trigonometrician knows that, if:

each is of the same frequency: $f_1 = f_2$ Hz 3

each is of the same amplitude: $V_1 = V_2$ volts 4

and they are 180^{o} out of phase: $\alpha = 180$ degrees 5

then: $y(t) = 0$ 6

A block diagram to represent eqn.(1) is suggested in Figure 1.

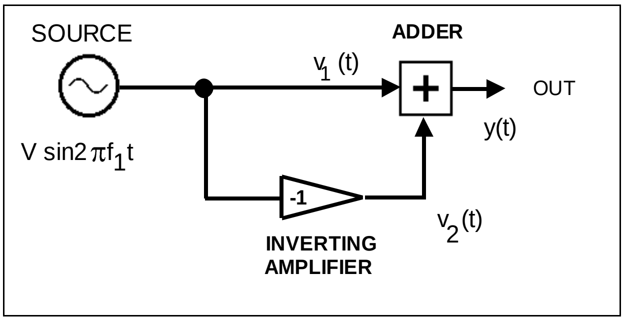


Figure 1: block diagram model of Equation 1

Note that we ensure the two signals are of the same frequency ($f_1 = f_2$) by obtaining them from the same source. The 180 degree phase change is achieved with an inverting amplifier, of unity gain.

In the block diagram of Figure 1 it is assumed, by convention, that the ADDER has unity gain between each input and the output. Thus the output is y(t) of eqn.(2).

This diagram appears to satisfy the requirements for obtaining a null at the output. Now see how we could model it with TIMS modules.

A suitable arrangement is illustrated in block diagram form in Figure 2.

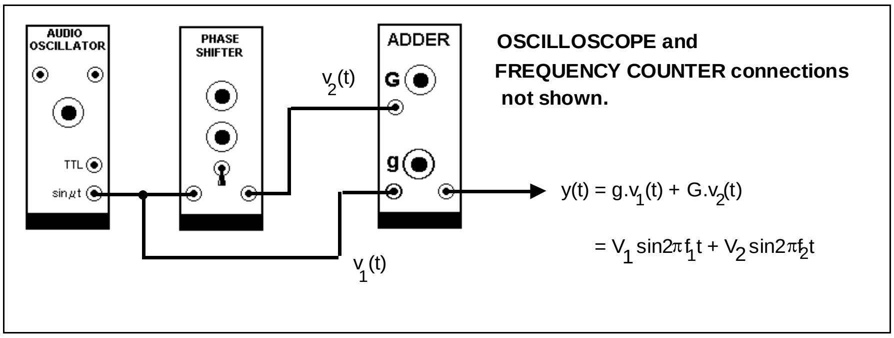


Figure 2: the TIMS model of Figure 1.

Before you build this model with TIMS modules let us consider the procedure you might follow in performing the experiment.

the ADDER

The annotation for the ADDER needs explanation. The symbol '**G**' near input **A** means the signal at this input will appear at the output, amplified by a factor '**G**'. Similar remarks apply to the input labelled '**g**'. Both '**G**' and '**g**' are adjustable by adjacent controls on the front panel of the ADDER. But note that, like the controls on all of the other TIMS modules, these controls are *not calibrated*. You must adjust these gains for a desired final result by measurement.

Thus the ADDER output is not identical with eqn.(2), but instead:

$$\text{ADDER output} = \mathbf{g}.v_1(t) + \mathbf{G}.v_2(t) \qquad \text{........ 7}$$

$$= V_1 \sin 2\pi f_1 t + V_2 \sin 2\pi f_2 t \qquad \text{........ 8}$$

conditions for a null

For a null at the output, sometimes referred to as a 'balance', one would be excused for thinking that:

if:

1) the PHASE SHIFTER is adjusted to introduce a difference of 180º between its input and output

and

2) the gains '**g**' and '**G**' are adjusted to equality

then

3) the amplitude of the output signal y(t) will be zero.

In practice the above procedure will almost certainly *not* result in zero output ! Here is the first *important observation* about the practical modelling of a theoretical concept.

In a practical system there are inevitably small impairments to be accounted for. For example, the gain through the PHASE SHIFTER is *approximately* unity, not exactly so. It would thus be pointless to set the gains '**g**' and '**G**' to be precisely equal. Likewise it would be a waste of time to use an expensive phase meter to set the PHASE SHIFTER to exactly 180º, since there are always small phase shifts not accounted for elsewhere in the model. See ***Q1, Tutorial Questions***, at the end of this experiment.

*These small impairments are **unknown**, but they are **stable**. Once compensated for they produce no further problems.*

So we do not make precise adjustments to modules, independently of the system into which they will be incorporated, and then patch them together and expect the system to behave. All adjustments are made *to the system as a whole* to bring about the desired end result.

The null at the output of the simple system of Figure 2 is achieved by adjusting the uncalibrated controls of the ADDER and of the PHASE SHIFTER. Although equations (3), (4), and (5) define the necessary conditions for a null, they do not give any guidance as to how to achieve these conditions.

more insight into the null

It is instructive to express eqn. (1) in phasor form. Refer to Figure 3.

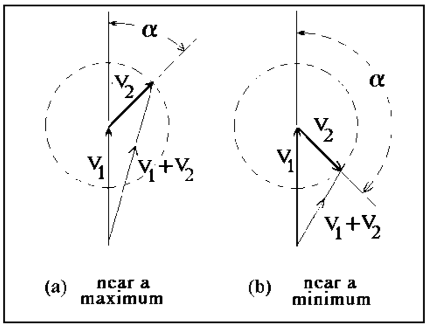


Figure 3: Equation (1) in phasor form

Figure 3 (a) and (b) shows the phasors V_1 and V_2 at two different angles α. It is clear that, to minimise the length of the resultant phasor ($V_1 + V_2$), the angle α in (b) needs to be increased by about 45°.

The resultant having reached a minimum, then V_2 must be increased to approach the magnitude of V_1 for an even smaller (finally zero) resultant.

We knew that already. What is clarified is the condition prior to the null being achieved. Note that, as angle α is rotated through a full 360°, the resultant ($V_1 + V_2$) goes through one minimum and one maximum (refer to the ***TIMS User Manual*** to see what sort of phase range is available from the PHASE SHIFTER).

What is also clear from the phasor diagram is that, when V_1 and V_2 differ by more than about 2:1 in magnitude, the minimum will be shallow, and the maximum broad and not pronounced [11].

Thus we can conclude that, unless the magnitudes V_1 and V_2 are already reasonably close, it may be difficult to find the null by rotating the phase control.

So, as a first step towards finding the null, it would be wise to set V_2 close to V_1. This will be done in the procedures detailed below.

Note that, for balance, it is the *ratio* of the magnitudes V_1 and V_2 , rather than their *absolute magnitudes*, which is of importance.

So we will consider V_1 of fixed magnitude (the reference), and make all adjustments to V_2.

This assumes V_1 is not of zero amplitude !

[11] fix V_1 as reference; mentally rotate the phasor for V_2. The dashed circle shows the locus of its extremity.

TIMS experiment procedures.

In each experiment the tasks '***T'*** you are expected to perform, and the questions '***Q***' you are expected to answer, are printed in italics and in slightly larger characters than the rest of the text.

In the early experiments there will a large list of tasks, each given in considerable detail. Later, you will not need such precise instructions, and only the major steps will be itemised. You are expected to become familiar with the capabilities of your oscilloscope, and especially with synchronization techniques.

Experiment

You are now ready to model eqn. (1). The modelling is explained step-by-step as a series of small tasks.

Take these tasks seriously, now and in later experiments, and TIMS will provide you with hours of stimulating experiences in telecommunications and beyond. The tasks are identified with a '***T***', are numbered sequentially, and should be performed in the order given.

T1 *both channels of the oscilloscope should be permanently connected to the matching coaxial connectors on the SCOPE SELECTOR. See the **TIMS User Manual** for details of this module.*

T2 *in this experiment you will be using three plug-in modules, namely: an AUDIO OSCILLATOR, a PHASE SHIFTER, and an ADDER. Obtain one each of these. Identify their various features as described in the **TIMS User Manual.***

Most modules can be controlled entirely from their front panels, but some have switches mounted on their circuit boards. Set these switches before plugging the modules into the TIMS SYSTEM UNIT; they will seldom require changing during the course of an experiment.

T3 *set the on-board range switch of the PHASE SHIFTER to 'LO'. Its circuitry is designed to give a wide phase shift in either the audio frequency range (LO), or the 100 kHz range (HI).*

Modules can be inserted into any one of the twelve available slots in the TIMS SYSTEM UNIT. Choose their locations to suit yourself. Typically one would try to match their relative locations as shown in the block diagram being modelled. Once plugged in, modules are in an operating condition.

T4 *plug the three modules into the TIMS SYSTEM UNIT.*

T5 *set the front panel switch of the FREQUENCY COUNTER to a GATE TIME of 1s. This is the most common selection for measuring frequency.*

When you become more familiar with TIMS you may choose to associate certain signals with particular patch lead colours. For the present, choose any colour which takes your fancy.

T6 *connect a patch lead from the lower yellow (analog) output of the AUDIO OSCILLATOR to the ANALOG input of the FREQUENCY COUNTER. The display will indicate the oscillator frequency f_1 in kilohertz (kHz).*

T7 *set the frequency f_1 with the knob on the front panel of the AUDIO OSCILLATOR, to approximately 1 kHz (any frequency would in fact be suitable for this experiment).*

T8 *connect a patch lead from the lower yellow (analog) output of the AUDIO OSCILLATOR to the 'ext. trig' [or 'ext. synch'] terminal of the oscilloscope. Make sure the oscilloscope controls are switched so as to accept this external trigger signal; use the automatic sweep mode if it is available.*

T9 *set the sweep speed of the oscilloscope to 0.5 ms/cm.*

T10 *patch a lead from the lower analog output of the AUDIO OSCILLATOR to the input of the PHASE SHIFTER.*

T11 *patch a lead from the output of the PHASE SHIFTER to the input **A** of the ADDER [12].*

T12 *patch a lead from the lower analog output of the AUDIO OSCILLATOR to the input **B** of the ADDER.*

T13 *patch a lead from the input **B** of the ADDER to CH2-A of the SCOPE SELECTOR module. Set the lower toggle switch of the SCOPE SELECTOR to UP.*

T14 *patch a lead from the input **A** of the ADDER to CH1-A of the SCOPE SELECTOR. Set the upper SCOPE SELECTOR toggle switch UP.*

T15 *patch a lead from the output of the ADDER to CH1-B of the SCOPE SELECTOR. This signal, y(t), will be examined later on.*

Your model should be the same as that shown in Figure 4 below, which is based on Figure 2. Note that in future experiments the format of Figure 2 will be used for TIMS models, rather than the more illustrative and informal style of Figure 4, which depicts the actual flexible patching leads.

You are now ready to set up some signal levels.

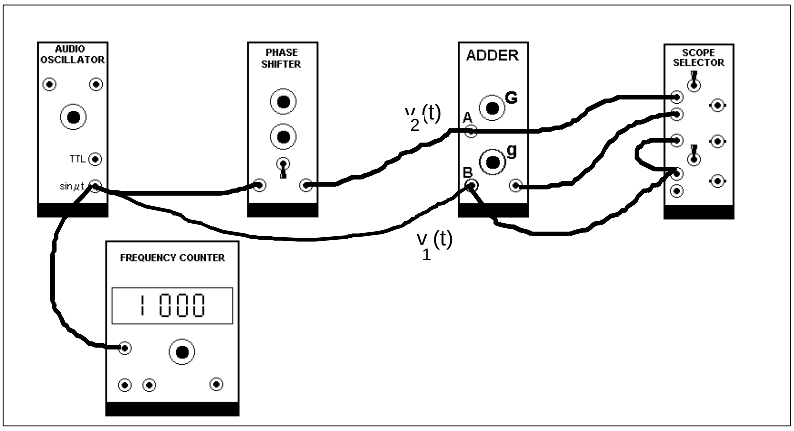


Figure 4: the TIMS model.

T16 *find the sinewave on CH1-A and, using the oscilloscope controls, place it in the upper half of the screen.*

12 the input is labelled 'A', and the gain is 'G'. This is often called 'the input G'; likewise 'input g'.

T17 *find the sinewave on CH2-A and, using the oscilloscope controls, place it in the lower half of the screen. This will display, throughout the experiment, a constant amplitude sine wave, and act as a monitor on the signal you are working with.*

Two signals will be displayed. These are the signals connected to the two ADDER *inputs*. One goes via the PHASE SHIFTER, which has a gain whose nominal value is unity; the other is a direct connection. They will be of the same nominal amplitude.

T18 *vary the COARSE control of the PHASE SHIFTER, and show that the relative phases of these two signals may be adjusted. Observe the effect of the* $\pm 180^0$ *toggle switch on the front panel of the PHASE SHIFTER.*

As part of the plan outlined previously it is now necessary to set the amplitudes of the two signals at the *output* of the ADDER to approximate equality.

Comparison of eqn. (1) with Figure 2 will show that the ADDER gain control **g** will adjust V_1, and **G** will adjust V_2.

You should set both V_1 and V_2, which are the *magnitudes* of the two signals at the ADDER *output*, at or near the TIMS ANALOG REFERENCE LEVEL, namely 4 volt peak-to-peak.

Now let us look at these two signals at the output of the ADDER.

T19 *switch the SCOPE SELECTOR from CH1-A to CH1-B. Channel 1 (upper trace) is now displaying the ADDER output.*

T20 *remove the patch cords from the **B** input of the ADDER. This sets the amplitude* V_1 *at the ADDER output to zero; it will not influence the adjustment of **G**.*

T21 *adjust the **G** gain control of the ADDER until the signal at the output of the ADDER, displayed on CH1-B of the oscilloscope, is about 4 volt peak-to-peak. This is* V_2.

T22 *remove the patch cord from the **A** input of the ADDER. This sets the* V_2 *output from the ADDER to zero, and so it will not influence the adjustment of **g**.*

T23 *replace the patch cords previously removed from the **B** input of the ADDER, thus restoring* V_1.

T24 *adjust the **g** gain control of the ADDER until the signal at the output of the ADDER, displayed on CH1-B of the oscilloscope, is about 4 volt peak-to-peak. This is* V_1.

T25 *replace the patch cords previously removed from the **A** input of the ADDER.*

Both signals (amplitudes V_1 and V_2) are now displayed on the upper half of the screen (CH1-B). Their individual amplitudes have been made approximately equal. Their *algebraic sum* may lie anywhere between zero and 8 volt peak-to-peak, depending on the value of the phase angle α. It is true that 8 volt peak-to-peak would be in excess of the TIMS ANALOG REFERENCE LEVEL, but it won`t overload the oscilloscope, and in any case will soon be reduced to a null.

Your task is to adjust the model for a null at the ADDER output, as displayed on CH1-B of the oscilloscope.

You may be inclined to fiddle, in a haphazard manner, with the few front panel controls available, and hope that before long a null will be achieved. You may be successful in a few moments, but this is unlikely. Such an approach is definitely not recommended if you wish to develop good experimental practices.

Instead, you are advised to remember the plan discussed above. This should lead you straight to the wanted result with confidence, and the *satisfaction that instant and certain success* can give.

There are only *three* conditions to be met, as defined by equations (3), (4), and (5).

- the *first* of these is already assured, since the two signals are coming from a common oscillator.
- the *second* is approximately met, since the gains '**g**' and '**G**' have been adjusted to make V_1 and V_2, at the ADDER *output*, about equal.
- the *third* is unknown, since the front panel control of the PHASE SHIFTER is not calibrated [13].

It would thus seem a good idea to start by adjusting the phase angle α. So:

T26 *set the FINE control of the PHASE SHIFTER to its central position.*

T27 *whilst watching the upper trace, y(t) on CH1-B, vary the COARSE control of the PHASE SHIFTER. Unless the system is at the null or maximum already, rotation in one direction will increase the amplitude, whilst in the other will reduce it. Continue in the direction which produces a decrease, until a minimum is reached. That is, when further rotation in the same direction changes the reduction to an increase. If such a minimum can not be found before the full travel of the COARSE control is reached, then reverse the front panel 180^O^ TOGGLE SWITCH, and repeat the procedure. Keep increasing the sensitivity of the oscilloscope CH1 amplifier, as necessary, to maintain a convenient display of y(t).*

Leave the PHASE SHIFTER controls in the position which gives the minimum.

T28 *now select the* ***G*** *control on the ADDER front panel to vary* V_2*, and rotate it in the direction which produces a deeper null. Since* V_1 *and* V_2 *have already been made almost equal, only a small change should be necessary.*

[13] TIMS philosophy is not to calibrate any controls. In this case it would not be practical, since the phase range of the PHASE SHIFTER varies with frequency.

T29 *repeating the previous two tasks a few times should further improve the depth of the null. As the null is approached, it will be found easier to use the* FINE *control of the PHASE SHIFTER. These adjustments (of amplitude and phase) are NOT interactive, so you should reach your final result after only a few such repetitions.*

Nulling of the two signals is complete !
You have achieved your first objective

You will note that it is not possible to achieve zero output from the ADDER. This never happens in a practical system. Although it *is* possible to reduce y(t) to zero, this cannot be observed, since it is masked by the inevitable system noise.

T30 *reverse the position of the PHASE SHIFTER toggle switch. Record the amplitude of y(t), which is now the absolute sum of V_1 PLUS V_2. Set this signal to fill the upper half of the screen. When the 180^0 switch is flipped back to the null condition, with the oscilloscope gain unchanged, the null signal which remains will appear to be 'almost zero'.*

signal-to-noise ratio

When y(t) is reduced in amplitude, by nulling to well below the TIMS ANALOG REFERENCE LEVEL, and the sensitivity of the oscilloscope is increased, the inevitable noise becomes visible. *Here noise is defined as anything we don`t want.*

The noise level will not be influenced by the phase cancellation process which operates on the test signal, so will remain to mask the moment when y(t) vanishes; see ***Q2***.

It will be at a level considered to be negligible in the TIMS environment - say less then 10 mV peak-to-peak. How many dB below reference level is this ?

Note that the nature of this noise can reveal many things. See ***Q3***.

achievements

Compared with some of the models you will be examining in later experiments you have just completed a very simple exercise. Yet many experimental techniques have been employed, and it is fruitful to consider some of these now, in case they have escaped your attention.

- to achieve the desired proportions of two signals V_1 and V_2 at the output of an ADDER it is necessary to measure first one signal, then the other. Thus it is necessary to remove the patch cord from one input whilst adjusting the output from the other. Turning the unwanted signal off with the front panel gain control is not a satisfactory method, since the original gain setting would then be lost.

- as the amplitude of the signal y(t) was reduced to a small value (relative to the remaining noise) it remained stationary on the screen. This was because the oscilloscope was triggering to a signal related in frequency (the same, in this case) and of *constant amplitude*, and was *not affected* by the nulling procedure. So the triggering circuits of the oscilloscope, once adjusted, remained adjusted.
- choice of the oscilloscope trigger signal is important. Since the oscilloscope remained synchronized, and a copy of y(t) remained on display (CH1) throughout the procedure, you could distinguish between the signal you were nulling and the accompanying noise.
- remember that the nulling procedure was focussed on the signal at the oscillator (fundamental) frequency. Depending on the nature of the remaining unwanted signals (noise) at the null condition, different conclusions can be reached.

  a) if the AUDIO OSCILLATOR had a significant amount of harmonic distortion, then the remaining 'noise' would be due to the presence of these harmonic components. It would be unlikely for them to be simultaneously nulled. The 'noise' would be stationary relative to the wanted signal (on CH1). The waveform of the 'noise' would provide a clue as to the order of the largest unwanted harmonic component (or components).

  b) if the remaining noise is entirely independent of the waveform of the signal on CH1, then one can make statements about the waveform purity of the AUDIO OSCILLATOR.

as time permits

At TRUNKS is a speech signal. You can identify it by examining each of the three TRUNKS outputs with your oscilloscope. You will notice that, during speech pauses, there remains a constant amplitude sinewave. This represents an interfering signal.

T31 *connect the speech signal at TRUNKS to the input of the HEADPHONE AMPLIFIER. Plug the headphones into the HEADPHONE AMPLIFIER, and listen to the speech. Notice that, no matter in which position the front panel switch labelled 'LPF Select' is switched, there is little change (if any at all) to the sound heard.*

There being no significant change to the sound means that the speech was already bandlimited to about 3 kHz, the LPF cutoff frequency, and that the interfering tone was within the same bandwidth. What would happen if this corrupted speech signal was used as the input to your model of Figure 2 ? Would it be possible to cancel out the interfering tone without losing the speech ?

T32 *connect the corrupted speech to your nulling model, and try to remove the tone from the speech. Report and explain results.*

Tutorial Questions

Q1 *refer to the phasor diagram of Figure 3. If the amplitudes of the phasors V_1 and V_2 were within 1% of each other, and the angle α within 1^o of 180^o, how would you describe the depth of null ? How would you describe the depth of null you achieved in the experiment ? You must be able to express the result numerically.*

Q2 *why was not the noise nulled at the same time as the 1 kHz test signal ?*

Q3 *describe a method (based on this experiment) which could be used to estimate the harmonic distortion in the output of an oscillator.*

Q4 *suppose you have set up the system of Figure 2, and the output has been successfully minimized. What might happen to this minimum if the frequency of the AUDIO OSCILLATOR was changed (say by 10%). Explain.*

Q5 *Figure 1 shows an INVERTING AMPLIFIER, but Figure 2 has a PHASE SHIFTER in its place. Could you have used a BUFFER AMPLIFIER (which inverts the polarity) instead of the PHASE SHIFTER ? Explain.*

TRUNKS

There should be a speech signal, corrupted by one or two tones, at TRUNKS. If you do not have a TRUNKS system you could generate this signal yourself with a SPEECH module, an AUDIO OSCILLATOR, and an ADDER.

2. DSBSC GENERATION

ACHIEVEMENTS: *definition and modelling of a double sideband suppressed carrier (DSBSC) signal; introduction to the MULTIPLIER, VCO, 60 kHz LPF, and TUNEABLE LPF modules; spectrum estimation; multipliers and modulators.*

PREREQUISITES: *completion of the experiment entitled '**Modelling an equation**' in this Volume.*

Preparation

This experiment will be your introduction to the MULTIPLIER and the double sideband suppressed carrier signal, or DSBSC. This modulated signal was probably not the first to appear in an historical context, but it is the easiest to generate.

You will learn that all of these modulated signals are derived from low frequency signals, or 'messages'. They reside in the frequency spectrum at some higher frequency, being placed there by being multiplied with a higher frequency signal, usually called 'the carrier' [14].

definition of a DSBSC

Consider two sinusoids, or cosinusoids, $\cos\mu t$ and $\cos\omega t$. A double sideband suppressed carrier signal, or DSBSC, is defined as their product, namely:

$$\text{DSBSC} = E.\cos\mu t \,.\, \cos\omega t \qquad \text{........ } 9$$

Generally, and in the context of this experiment, it is understood that::

$$\omega >> \mu \qquad \text{........ } 10$$

Equation (3) can be expanded to give:

$$\cos\mu t \cdot \cos\omega t = (E/2)\cos(\omega - \mu)t + (E/2)\cos(\omega + \mu)t \qquad \text{...... } 11$$

Equation 3 shows that the product is represented by two new signals, one on the sum frequency $(\omega + \mu)$, and one on the difference frequency $(\omega - \mu)$ - see Figure 1.

Remembering the inequality of eqn. (2) the two new components are located close to the frequency ω rad/s, one just below, and the other just above it. These are referred to as the *lower* and *upper* sidebands [15] respectively.

---

[14] but remember whilst these *low* and *high* qualifiers reflect common practice, they are not mandatory.

[15] when, as here, there is only one component either side of the carrier, they are better described as *side frequencies*. With a more complex message there are many components either side of the carrier, from whence comes the term *sidebands*.

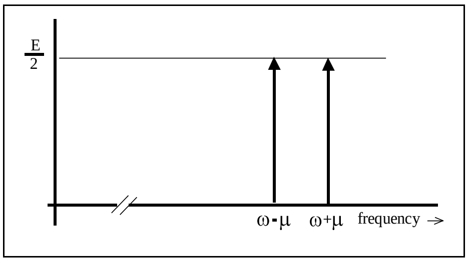


Figure 1: spectral components

These two components were derived from a 'carrier' term on ω rad/s, and a message on μ rad/s. Because there is no term at carrier frequency in the product signal it is described as a double sideband *suppressed* carrier (DSBSC) signal.

The term 'carrier' comes from the context of 'double sideband amplitude modulation' (commonly abbreviated to just AM).

AM is introduced in a later experiment (although, historically, AM preceded DSBSC).

The time domain appearance of a DSBSC (eqn. 1) in a text book is generally as shown in Figure 2.

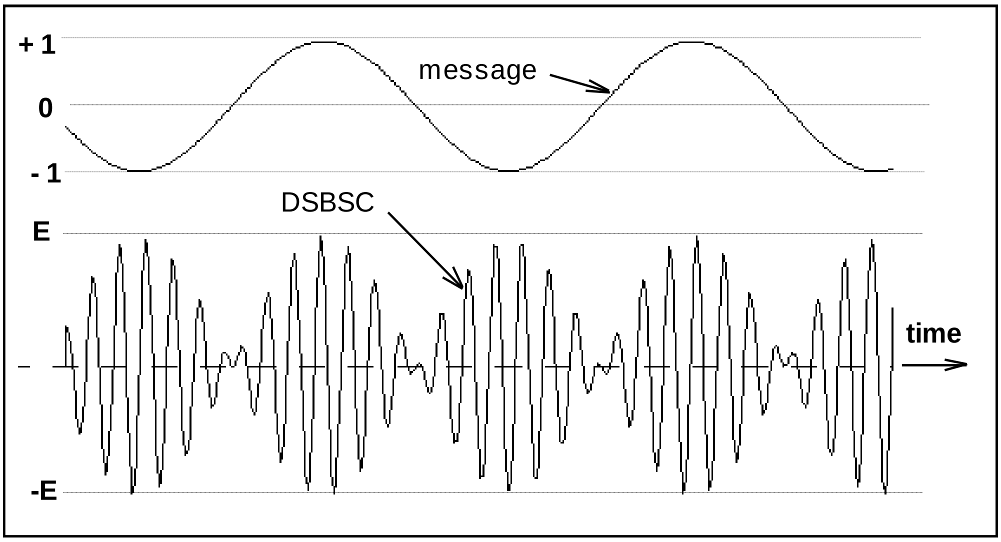


Figure 2: eqn.(1) - a DSBSC - seen in the time domain

Notice the waveform of the DSBSC in Figure 2, especially near the times when the message amplitude is zero. The fine detail differs from period to period of the message. This is because the ratio of the two frequencies μ and ω has been made non-integral.

Although the message and the carrier are periodic waveforms (sinusoids), the DSBSC itself need not necessarily be periodic.

block diagram

A block diagram, showing how eqn. (1) could be modelled with hardware, is shown in Figure 3 below.

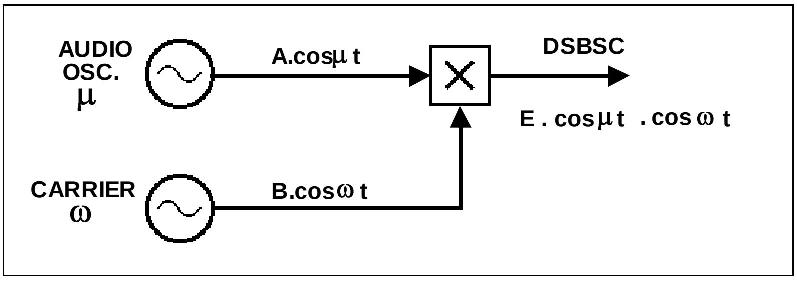


Figure 3: block diagram to generate eqn. (1) with hardware.

viewing envelopes

This is the first experiment dealing with a narrow band signal. Nearly all modulated signals in communications are narrow band. The definition of 'narrow band' has already been discussed in the chapter *Introduction to Modelling with TIMS*.

You will have seen pictures of DSB or DSBSC signals (and amplitude modulation - AM) in your text book, and probably have a good idea of what is meant by their envelopes [16]. You will only be able to reproduce the text book figures if the oscilloscope is set appropriately - especially with regard to the method of its synchronization. Any other methods of setting up will still be displaying the same signal, but not in the familiar form shown in text books. How is the 'correct method' of synchronization defined ?

With narrow-band signals, and particularly of the type to be examined in this and the modulation experiments to follow, the following steps are recommended:

1) use a single tone for the message, say 1 kHz.
2) synchronize the oscilloscope to the message generator, which is of fixed amplitude, using the 'ext trig.' facility.
3) set the sweep speed so as to display one or two periods of this message on one channel of the oscilloscope.
4) display the modulated signal on another channel of the oscilloscope.

With the recommended scheme the envelope will be stationary on the screen. In all but the most special cases the actual modulated waveform itself will not be stationary - since successive sweeps will show it in slightly different positions. So the display *within* the envelope - the modulated signal - will be 'filled in', as in Figure 4, rather than showing the detail of Figure 2.

[16] there are later experiments addressed specifically to envelopes, namely those entitled ***Envelopes***, and ***Envelope Recovery***.

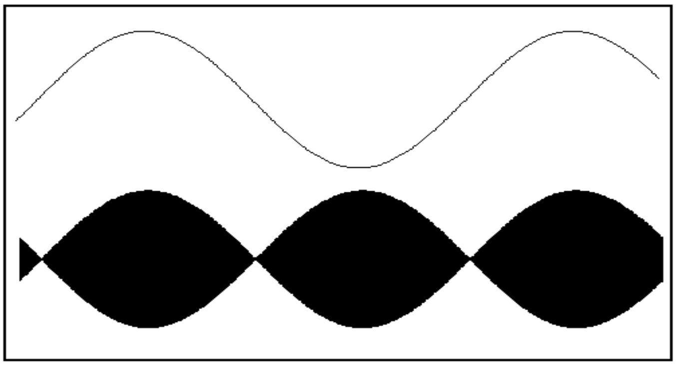


Figure 4: typical display of a DSBSC, with the message from which it was derived, as seen on an oscilloscope. Compare with Figure 2.

multi-tone message

The DSBSC has been defined in eqn. (1), with the message identified as the low frequency term. Thus:

$$\text{message} = \cos\mu t \qquad \text{........ 12}$$

For the case of a multi-tone message, m(t), where:

$$m(t) = \sum_{i=1}^{n} a_i \cos \mu_i t \qquad \text{........ 13}$$

then the corresponding DSBSC signal consists of a *band* of frequencies below ω, and a *band* of frequencies above ω. Each of these bands is of width equal to the bandwidth of m(t).

The individual spectral components in these sidebands are often called sidefrequencies.

If the frequency of each term in the expansion is expressed in terms of its difference from ω, and the terms are grouped in pairs of sum and difference frequencies, then there will be 'n' terms of the form of the right hand side of eqn. (3).

Note it is assumed here that there is no DC term in m(t). The presence of a DC term in m(t) will result in a term at ω in the DSB signal; that is, a term at 'carrier' frequency. It will no longer be a double sideband *suppressed* carrier signal. A special case of a DSB with a significant term at carrier frequency is an *amplitude modulated* signal, which will be examined in an experiment to follow.

A more general definition still, of a DSBSC, would be:

$$\text{DSBSC} = \text{E.m(t).}\cos\omega t \qquad \text{........ 14}$$

where m(t) is any (low frequency) message. By convention m(t) is generally understood to have a peak amplitude of unity (and typically no DC component).

linear modulation

The DSBSC is a member of a class known as *linear modulated* signals. Here the spectrum of the modulated signal, when the message has two or more components, is the sum of the spectral components which each message component would have produced if present alone.

For the case of *non-linear modulated* signals, on the other hand, this linear addition does not take place. In these cases the whole is more than the sum of the parts. A frequency modulated (FM) signal is an example. These signals are first examined in the chapter entitled *Analysis of the FM spectrum*, within *Volume A2 - Further & Advanced Analog Experiments*, and subsequent experiments of that Volume.

spectrum analysis

In the experiment entitled *Spectrum analysis - the WAVE ANALYSER*, within *Volume A2 - Further & Advanced Analog Experiments*, you will model a *WAVE ANALYSER*. As part of that experiment you will re-examine the DSBSC spectrum, paying particular attention to its spectrum.

Experiment

the MULTIPLIER

This is your introduction to the MULTIPLIER module.

Please read the section in the chapter of this Volume entitled *Introduction to modelling with TIMS* headed *multipliers and modulators*. Particularly note the comments on DC off-sets.

preparing the model

Figure 3 shows a block diagram of a system suitable for generating DSBSC derived from a single tone message.

Figure 5 shows how to model this block diagram with TIMS.

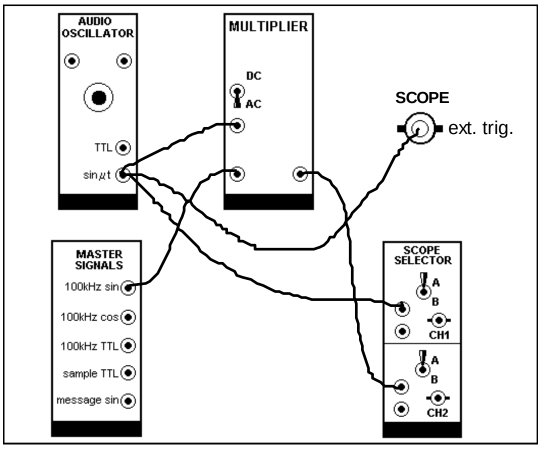


Figure 5: pictorial of block diagram of Figure 3

The signal A.cosμt, of fixed amplitude **A**, from the AUDIO OSCILLATOR, represents the single tone message. A signal of fixed amplitude from this oscillator is used to synchronize the oscilloscope.

The signal B.cosωt, of fixed amplitude **B** and frequency exactly 100 kHz, comes from the MASTER SIGNALS panel. This is the TIMS high frequency, or radio, signal. Text books will refer to it as the 'carrier signal'.

The amplitudes **A** and **B** are nominally equal, being from TIMS signal sources. They are suitable as inputs to the MULTIPLIER, being at the TIMS ANALOG REFERENCE LEVEL. The output from the MULTIPLIER will also be, by design of the internal circuitry, at this nominal level. There is no need for any amplitude adjustment. It is a very simple model.

T1 *patch up the arrangement of Figure 5. Notice that the oscilloscope is triggered by the message,* ***not*** *the DSBSC itself (nor, for that matter, by the carrier).*

T2 *use the FREQUENCY COUNTER to set the AUDIO OSCILLATOR to about 1 kHz*

Figure 2 shows the way most text books would illustrate a DSBSC signal of this type. But the display you have in front of you is more likely to be similar to that of Figure 4.

signal amplitude.

T3 *measure and record the amplitudes* ***A*** *and* ***B*** *of the message and carrier signals at the inputs to the MULTIPLIER.*

The output of this arrangement is a DSBSC signal, and is given by:

$$\text{DSBSC} = k\ A.\cos\mu t\ B.\cos\omega t \qquad \text{...... } 15$$

The peak-to-peak amplitude of the display is:

$$\text{peak-to-peak} = 2\,k\,A\,B \text{ volts} \qquad \text{......}\ 16$$

Here 'k' is a scaling factor, a property of the MULTIPLIER. One of the purposes of this experiment is to determine the magnitude of this parameter.

Now:

T4 *measure the peak-to-peak amplitude of the DSBSC*

Since you have measured both **A** and **B** already, you have now obtained the magnitude of the MULTIPLIER scale factor 'k'; thus:

$$k = (\text{dsbsc peak-to-peak}) / (2\,A\,B) \qquad \text{......}\ 17$$

Note that 'k' is not a dimensionless quantity.

fine detail in the time domain

The oscilloscope display will not in general show the fine detail inside the DSBSC, yet many textbooks will do so, as in Figure 2. Figure 2 would be displayed by a single sweep across the screen. The normal laboratory oscilloscope cannot retain and display the picture from a single sweep [17]. Subsequent sweeps will all be slightly different, and will not coincide when superimposed.

To make consecutive sweeps identical, and thus to display the DSBSC as depicted in Figure 2, it is necessary that ‘μ’ be a sub-multiple of ‘ω’. This special condition can be arranged with TIMS by choosing the '2 kHz MESSAGE' sinusoid from the fixed MASTER SIGNALS module. The frequency of this signal is actually 100/48 kHz (approximately 2.08 kHz), an exact sub-multiple of the carrier frequency. Under these special conditions the fine detail of the DSBSC can be observed.

T5 *obtain a display of the DSBSC similar to that of Figure 2. A sweep speed of, say, 50μs/cm is a good starting point.*

overload

When designing an analog system signal overload must be avoided at all times. Analog circuits are expected to operate in a linear manner, in order to reduce the chance of the generation of new frequencies. This would signify non-linear operation.

A multiplier is *intended* to generate new frequencies. In this sense it is a non-linear device. Yet it should only produce those new frequencies which are *wanted* - any other frequencies are deemed *unwanted*.

A quick test for unintended (non-linear) operation is to use it to generate a signal with a known shape -a DSBSC signal is just such a signal. Presumably so far your MULTIPLIER module has been behaving ‘linearly’.

[17] but note that, since the oscilloscope is synchronized to the message, the envelope of the DSBSC remains in a fixed relative position over consecutive sweeps. It is the infill - the actual DSBSC itself - which is slightly different each sweep.

T6 *insert a BUFFER AMPLIFIER in one or other of the paths to the MULTIPLIER, and increase the input amplitude of this signal until overload occurs. Sketch and describe what you see.*

bandwidth

Equation (3) shows that the DSBSC signal consists of two components in the frequency domain, spaced above and below ω by μ rad/s.

With the TIMS BASIC SET of modules, and a DSBSC based on a 100 kHz carrier, you can make an indirect check on the truth of this statement. Attempting to pass the DSBSC through a 60 kHz LOWPASS FILTER will result in no output, evidence that the statement has some truth in it - all components must be above 60 kHz.

A convincing proof can be made with the 100 kHz CHANNEL FILTERS module [18]. Passage through any of these filters will result in no change to the display (see *alternative spectrum check* later in this experiment).

Using only the resources of the TIMS BASIC SET of modules a convincing proof is available if the carrier frequency is changed to, say, 10 kHz. This signal is available from the analog output of the VCO, and the test setup is illustrated in Figure 6 below. Lowering the carrier frequency puts the DSBSC in the range of the TUNEABLE LPF.

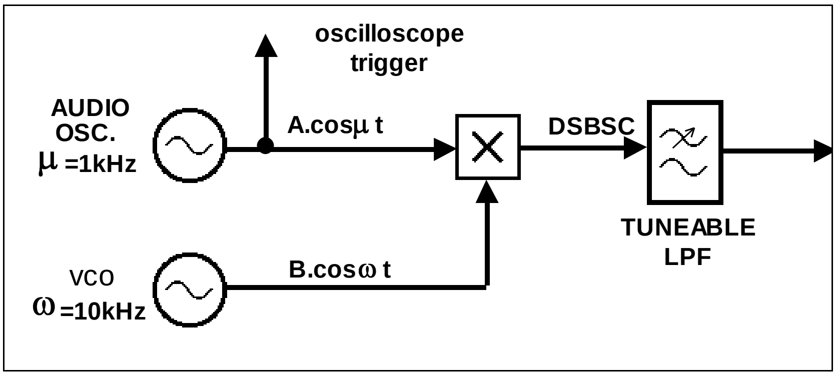


Figure 6: checking the spectrum of a DSBSC signal

T7 *read about the VCO module in the* ***TIMS User Manual.*** *Before plugging the VCO in to the TIMS SYSTEM UNIT set the on-board switch to VCO. Set the front panel frequency range selection switch to 'LO'.*

T8 *read about the TUNEABLE LPF in the* ***TIMS User Manual*** *and the* ***Appendix A*** *to this text.*

T9 *set up an arrangement to check out the TUNEABLE LPF module. Use the VCO as a source of sinewave input signal. Synchronize the oscilloscope to this signal. Observe input to, and output from, the TUNEABLE LPF.*

T10 *set the front panel GAIN control of the TUNEABLE LPF so that the gain through the filter is unity.*

18 this is a TIMS ADVANCED MODULE.

T11 *confirm the relationship between VCO frequency and filter cutoff frequency (refer to the* **TIMS User Manual** *for full details, or the Appendix to this Experiment for abridged details).*

T12 *set up the arrangement of Figure 6. Your model should look something like that of Figure 7, where the arrangement is shown modelled by TIMS.*

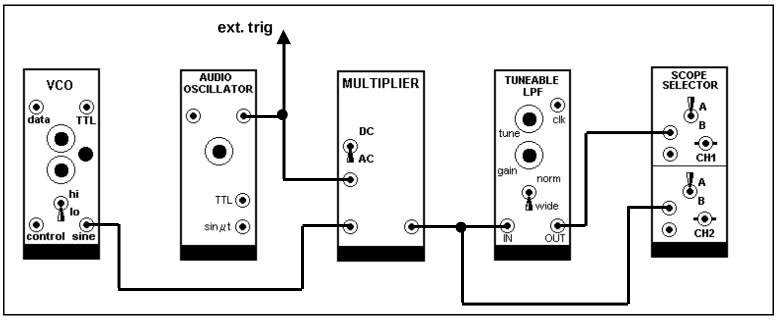


Figure 7: TIMS model of Figure 6

T13 *adjust the VCO frequency to about 10 kHz*

T14 *set the AUDIO OSCILLATOR to about 1 kHz.*

T15 *confirm that the output from the MULTIPLIER looks like Figures 2 and/or 4.*

Analysis predicts that the DSBSC is centred on 10 kHz, with lower and upper sidefrequencies at 9.0 kHz and 11.0 kHz respectively. Both sidefrequencies should fit well within the passband of the TUNEABLE LPF, when it is tuned to its widest passband, and so the shape of the DSBSC should not be altered.

T16 *set the front panel toggle switch on the TUNEABLE LPF to* WIDE*, and the front panel* TUNE *knob fully clockwise. This should put the passband edge above 10 kHz. The passband edge (sometimes called the 'corner frequency') of the filter can be determined by connecting the output from the TTL* CLK *socket to the FREQUENCY COUNTER. It is given by dividing the counter readout by 360 (in the 'NORMAL' mode the dividing factor is 880).*

T17 *note that the passband* GAIN *of the TUNEABLE LPF is adjustable from the front panel. Adjust it until the output has a similar amplitude to the DSBSC from the MULTIPLIER (it will have the same shape). Record the width of the passband of the TUNEABLE LPF under these conditions.*

Assuming the last Task was performed successfully this confirms that the DSBSC lies below the passband edge of the TUNEABLE LPF at its widest. You will now use the TUNEABLE LPF to determine the sideband locations. That this should be possible is confirmed by Figure 8 below.

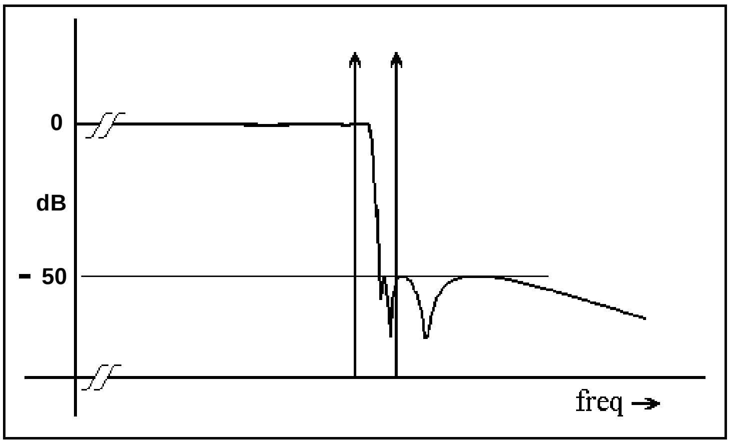


Figure 8: the amplitude response of the TUNEABLE LPF superimposed on the DSBSC spectrum.

Figure 8 shows the amplitude response of the TUNEABLE LPF superimposed on the DSBSC, when based on a 1 kHz message. The drawing is approximately to scale. It is clear that, with the filter tuned as shown (passband edge just above the lower sidefrequency), it *is* possible to attenuate the upper sideband by 50 dB and retain the lower sideband effectively unchanged.

T18 *make a sketch to explain the meaning of the* <u>*transition bandwidth*</u> *of a lowpass filter. You should measure the transition bandwidth of your TUNEABLE LPF, or instead accept the value given in* ***Appendix A*** *to this text.*

T19 *lower the filter passband edge until there is a just-noticeable change to the DSBSC output. Record the filter passband edge as f_A. You have located the upper edge of the DSBSC at $(\omega + \mu)$ rad/s.*

T20 *lower the filter passband edge further until there is only a sinewave output. You have isolated the component on $(\omega - \mu)$ rad/s. Lower the filter passband edge still further until the amplitude of this sinewave just starts to reduce. Record the filter passband edge as f_B.*

T21 *again lower the filter passband edge, just enough so that there is no significant output. Record the filter passband edge as f_C*

T22 *from a knowledge of the filter transition band ratio, and the measurements f_A and f_B , estimate the location of the two sidebands and compare with expectations. You could use f_C as a cross-check.*

alternative spectrum check

If you have a 100kHz CHANNEL FILTERS module, or from a SPEECH module, then, knowing the filter bandwidth, it can be used to verify the theoretical estimate of the DSBSC bandwidth.

speech as the message

If you have speech available at TRUNKS you might like to observe the appearance of the DSBSC signal in the time domain.

Figure 9 is a snap-shot of what you might see.

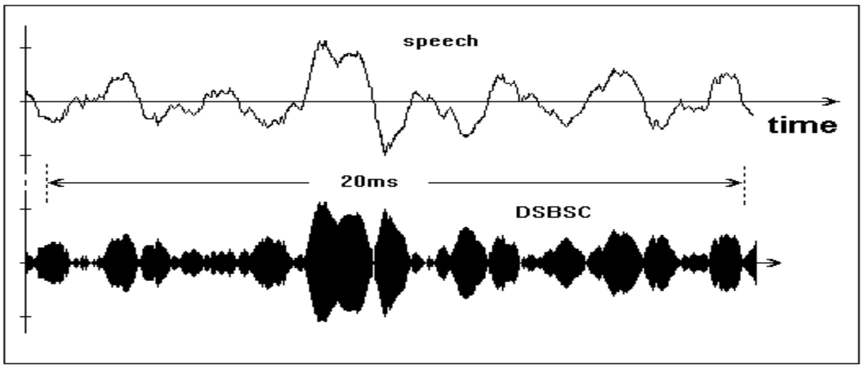


Figure 9: speech derived DSBSC

Tutorial Questions

Q1 *in TIMS the parameter 'k' has been set so that the product of two sinewaves, each at the TIMS ANALOG REFERENCE LEVEL, will give a MULTIPLIER peak-to-peak output amplitude also at the TIMS ANALOG REFERENCE LEVEL. Knowing this, predict the expected magnitude of 'k'*

Q2 *how would you answer the question 'what is the frequency of the signal* $y(t) = E.\cos\mu t.\cos\omega t$*' ?*

Q3 *what would the FREQUENCY COUNTER read if connected to the signal* $y(t) = E.\cos\mu t.\cos\omega t$ *?*

Q4 *is a DSBSC signal periodic ?*

Q6 *carry out the trigonometry to obtain the spectrum of a DSBSC signal when the message consists of three tones, namely:*

$$\text{message} = A_1.\cos\mu_1 t + A_2.\cos\mu_2 t + A_3 \cos\mu_3 t$$

Show that it is the linear sum of three DSBSC, one for each of the individual message components.

Q6 *the DSBSC definition of eqn. (1) carried the understanding that the message frequency* μ *should be very much less than the carrier frequency* ω*. Why was this ? Was it strictly necessary ? You will have an opportunity to consider this in more detail in the experiment entitled* ***Envelopes*** *(within Volume A2 - Further & Advanced Analog Experiments).*

TRUNKS

If you do not have a TRUNKS system you could obtain a speech signal from a SPEECH module.

Appendix

TUNEABLE LPF tuning information

Filter cutoff frequency is given by:

NORM range: clk / 880

WIDE range: clk / 360

See the ***TIMS User Manual*** for full details.

3. AMPLITUDE MODULATION

ACHIEVEMENTS: *modelling of an amplitude modulated (AM) signal; method of setting and measuring the depth of modulation; waveforms and spectra; trapezoidal display.*

PREREQUISITES: *a knowledge of DSBSC generation. Thus completion of the experiment entitled* **DSBSC generation** *would be an advantage.*

Preparation

In the early days of wireless, communication was carried out by telegraphy, the radiated signal being an interrupted radio wave. Later, the amplitude of this wave was varied in sympathy with (modulated by) a speech message (rather than on/off by a telegraph key), and the message was recovered from the envelope of the received signal. The radio wave was called a 'carrier', since it was seen to carry the speech information with it. The process and the signal was called *amplitude modulation*, or 'AM' for short.

In the context of radio *communications*, near the end of the 20th century, few modulated signals contain a significant component at 'carrier' frequency. However, despite the fact that a carrier is not radiated, the need for such a signal at the transmitter (where the modulated signal is generated), and also at the receiver, remains fundamental to the modulation and demodulation process respectively. The use of the term 'carrier' to describe this signal has continued to the present day.

As distinct from radio communications, present day radio *broadcasting* transmissions do have a carrier. By transmitting this carrier the design of the demodulator, at the receiver, is greatly simplified, and this allows significant cost savings.

The most common method of AM generation uses a 'class C modulated amplifier'; such an amplifier is not available in the BASIC TIMS set of modules. It is well documented in text books. This is a 'high level' method of generation, in that the AM signal is generated at a power level ready for radiation. It is still in use in broadcasting stations around the world, ranging in powers from a few tens of watts to many megawatts.

Unfortunately, text books which describe the operation of the class C modulated amplifier tend to associate properties of this particular method of generation with those of AM, and AM generators, in general. This gives rise to many misconceptions. The worst of these is the belief that it is impossible to generate an AM signal with a depth of modulation exceeding 100% without giving rise to serious RF distortion.

You will see in this experiment, and in others to follow, that there is no problem in generating an AM signal with a depth of modulation exceeding 100%, and without any RF distortion whatsoever.

But we are getting ahead of ourselves, as we have not yet even defined what AM is !

theory

The amplitude modulated signal is defined as:

$$AM = E\,(1 + m.\cos\mu t)\,\cos\omega t \qquad \text{........ } 18$$

$$= A\,(1 + m.\cos\mu t)\,.\,B\,\cos\omega t \qquad \text{........ } 19$$

$$= [\text{low frequency term } a(t)] \times [\text{high frequency term } c(t)] \qquad \text{........ } 20$$

Here:

'**E**' is the AM signal amplitude from eqn. (1). For modelling convenience eqn. (1) has been written into two parts in eqn. (2), where (A.B) = E.

'**m**' is a constant, which, as you will soon see, defines the 'depth of modulation'. Typically m < 1. Depth of modulation, expressed as a percentage, is 100.m. There is no inherent restriction upon the size of 'm' in eqn. (1). This point will be discussed later.

'**μ**' and '**ω**' are angular frequencies in rad/s, where $\mu/(2.\pi)$ is a low, or message frequency, say in the range 300 Hz to 3000 Hz; and $\omega/(2.\pi)$ is a radio, or relatively high, 'carrier' frequency. In TIMS the carrier frequency is generally 100 kHz.

Notice that the term a(t) in eqn. (3) contains both a DC component and an AC component. As will be seen, it is the DC component which gives rise to the term at ω - the 'carrier' - in the AM signal. The AC term 'm.cosμt' is generally thought of as the message, and is sometimes written as m(t). But strictly speaking, to be compatible with other mathematical derivations, the whole of the low frequency term a(t) should be considered the message.

Thus:

$$a(t) = DC + m(t) \qquad \text{........ } 21$$

Figure 1 below illustrates what the oscilloscope will show if displaying the AM signal.

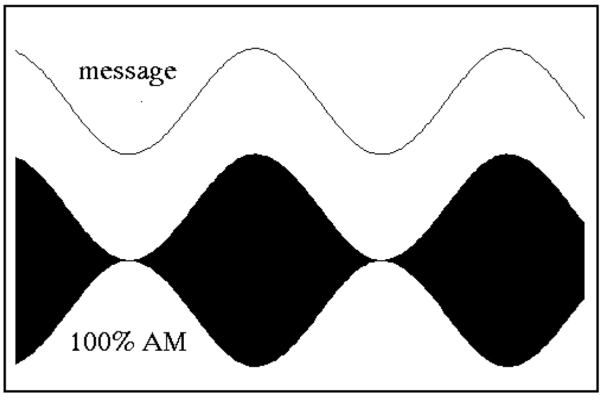


Figure 1 - AM, with m = 1, as seen on the oscilloscope

A block diagram representation of eqn. (2) is shown in Figure 2 below.

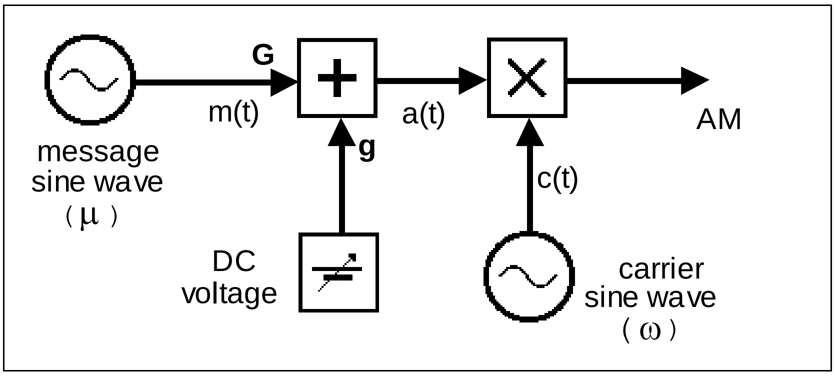


Figure 2: generation of equation 2

For the first part of the experiment you will model eqn. (2) by the arrangement of Figure 2. The depth of modulation will be set to exactly 100% (m = 1). You will gain an appreciation of the meaning of ‘depth of modulation’, and you will learn how to set other values of ‘m’, including cases where m > 1.

The signals in eqn. (2) are expressed as voltages in the time domain. You will model them in two parts, as written in eqn. (3).

depth of modulation

100% amplitude modulation is defined as the condition when m = 1. Just what this means will soon become apparent. It requires that the amplitude of the DC (= A) part of a(t) is equal to the amplitude of the AC part (= A.m). This means that their ratio is unity at the *output* of the ADDER, which forces ‘m’ to a magnitude of exactly unity.

> *By aiming for a ratio of unity it is thus not necessary to know the absolute magnitude of A at all.*

measurement of ‘m’

The magnitude of ‘m’ can be measured directly from the AM display itself.

Thus:

$$m = \frac{P - Q}{P + Q} \qquad \text{........ 22}$$

where P and Q are as defined in Figure 3.

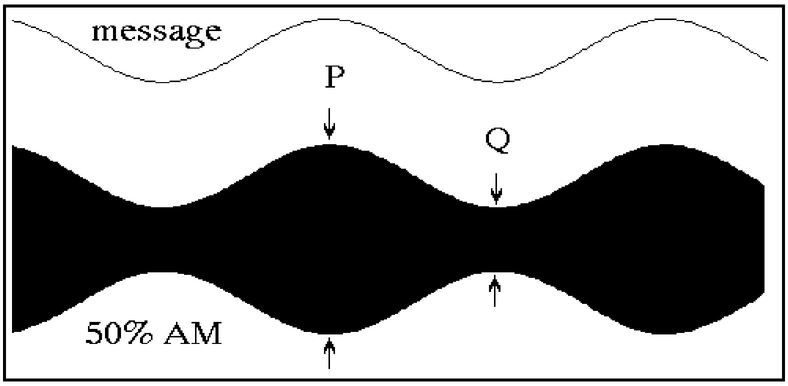


Figure 3: the oscilloscope display for the case m = 0.5

spectrum

Analysis shows that the sidebands of the AM, when derived from a message of frequency μ rad/s, are located either side of the carrier frequency, spaced from it by μ rad/s.

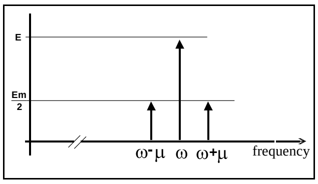


Figure 4: AM spectrum

You can see this by expanding eqn. (2). The spectrum of an AM signal is illustrated in Figure 4 (for the case m = 0.75). The spectrum of the DSBSC alone was confirmed in the experiment entitled *DSBSC generation*. You can repeat this measurement for the AM signal.

As the analysis predicts, even when m > 1, there is no widening of the spectrum.

This assumes linear operation; that is, that there is *no hardware overload.*

other message shapes.

Provided m ≤ 1 the envelope of the AM will always be a faithful copy of the message. For the generation method of Figure 2 the requirement is that:

> *the peak amplitude of the AC component must not exceed the magnitude of the DC, measured at the ADDER output*

As an example of an AM signal derived from speech, Figure 5 shows a snap-shot of an AM signal, and separately the speech signal.

There are no amplitude scales shown, but you should be able to deduce the depth of modulation [19] by inspection.

19 that is, the *peak* depth

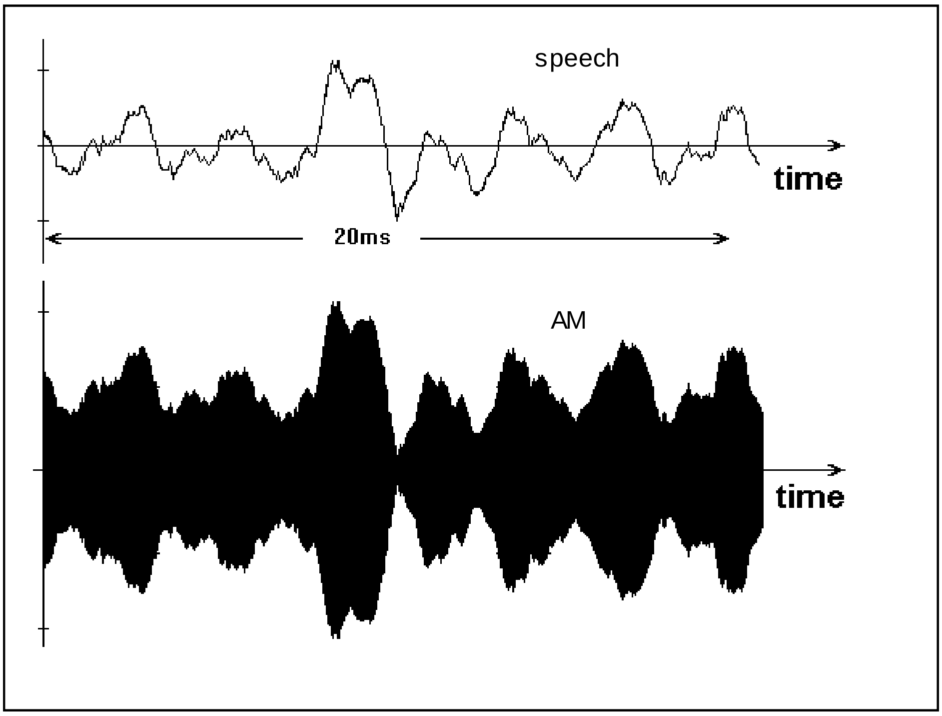


Figure 5: AM derived from speech.

other generation methods

There are many methods of generating AM, and this experiment explores only one of them. Another method, which introduces more variables into the model, is explored in the experiment entitled *Amplitude modulation - method 2*, to be found in *Volume A2 - Further & Advanced Analog Experiments*.

It is strongly suggested that you examine your text book for other methods.

Practical circuitry is more likely to use a *modulator*, rather than the more idealised *multiplier*. These two terms are introduced in the Chapter of this Volume entitled *Introduction to modelling with TIMS*, in the section entitled *multipliers and modulators.*

Experiment

aligning the model

the low frequency term a(t)

To generate a voltage defined by eqn. (2) you need first to generate the term a(t).

$$a(t) = A.(1 + m.\cos\mu t) \qquad \text{........ } 23$$

Note that this is the addition of two parts, a DC term and an AC term. Each part may be of any convenient amplitude at the *input* to an ADDER.

The DC term comes from the VARIABLE DC module, and will be adjusted to the amplitude 'A' at the *output* of the ADDER.

The AC term m(t) will come from an AUDIO OSCILLATOR, and will be adjusted to the amplitude 'A.m' at the *output* of the ADDER.

the carrier supply c(t)

The 100 kHz carrier c(t) comes from the MASTER SIGNALS module.

$$c(t) = B.\cos\omega t \qquad \text{........ 24}$$

The block diagram of Figure 2, which models the AM equation, is shown modelled by TIMS in Figure 6 below.

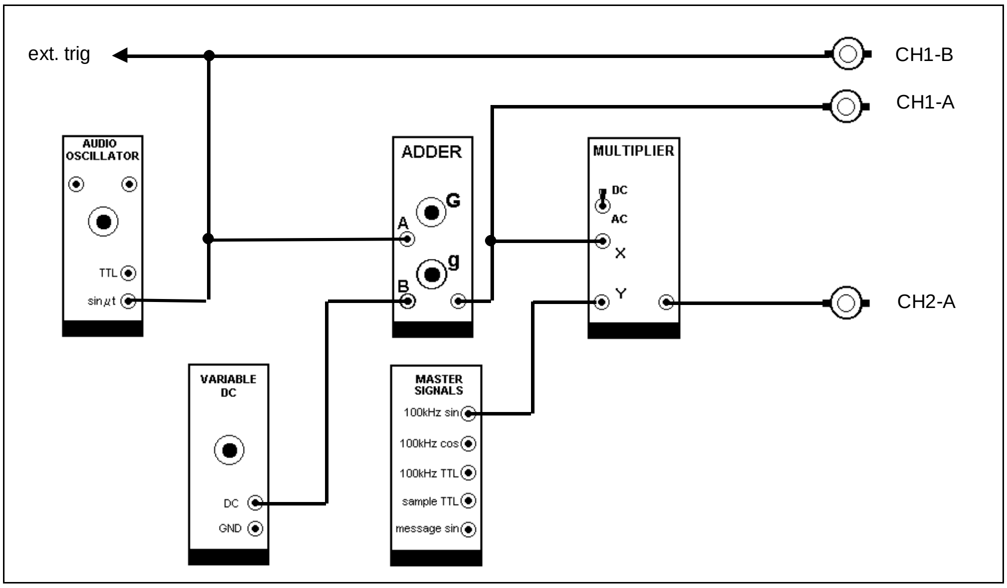


Figure 6: the TIMS model of the block diagram of Figure 2

To build the model:

T1 *first patch up according to Figure 6, but omit the input **X** and **Y** connections to the MULTIPLIER. Connect to the two oscilloscope channels using the SCOPE SELECTOR, as shown.*

T2 *use the FREQUENCY COUNTER to set the AUDIO OSCILLATOR to about 1 kHz.*

T3 *switch the SCOPE SELECTOR to CH1-B, and look at the message from the AUDIO OSCILLATOR. Adjust the oscilloscope to display two or three periods of the sine wave in the top half of the screen.*

Now start adjustments by setting up a(t), as defined by eqn. (4), and with m = 1.

T4 *turn both **g** and **G** fully anti-clockwise. This removes both the DC and the AC parts of the message from the output of the ADDER.*

T5 *switch the scope selector to CH1-A. This is the ADDER output. Switch the oscilloscope amplifier to respond to DC if not already so set, and the sensitivity to about 0.5 volt/cm. Locate the trace on a convenient grid line towards the bottom of the screen. Call this the* ***zero reference grid line****.*

T6 *turn the front panel control on the VARIABLE DC module almost fully anti-clockwise (not critical). This will provide an output voltage of about minus 2 volts. The ADDER will reverse its polarity, and adjust its amplitude using the '****g****' gain control.*

T7 *whilst noting the oscilloscope reading on CH1-A, rotate the gain '****g****' of the ADDER clockwise to adjust the DC term at the output of the ADDER to exactly 2 cm above the previously set zero reference line. This is 'A' volts.*

You have now set the magnitude of the DC part of the message to a known amount. This is *about* 1 volt, but *exactly* 2 cm, on the oscilloscope screen. You must now make the AC part of the message equal to this, so that the *ratio* **Am/A** will be unity. This is easy:

T8 *whilst watching the oscilloscope trace of CH1-A rotate the ADDER gain control '****G****' clockwise. Superimposed on the DC output from the ADDER will appear the message sinewave. Adjust the gain* ***G*** *until the lower crests of the sinewave are EXACTLY coincident with the previously selected* ***zero reference grid line****.*

The sine wave will be centred exactly **A** volts above the previously-chosen zero reference, and so its amplitude is **A**.

Now the DC and AC, each at the ADDER output, are of exactly the same amplitude **A**. Thus:

$$A = A.m \quad \text{(for } m = 1) \qquad \text{........ } 25$$

and so:

$$m = 1 \qquad \text{........ } 26$$

You have now modelled A.(1 + m.cosμt), with m = 1. This is connected to one input of the MULTIPLIER, as required by eqn. (2).

T9 *connect the output of the ADDER to input* ***X*** *of the MULTIPLIER. Make sure the MULTIPLIER is switched to accept DC.*

Now prepare the carrier signal:

$$c(t) = B.\cos\omega t \qquad \text{........ } 27$$

T10 *connect a 100 kHz analog signal from the MASTER SIGNALS module to input* ***Y*** *of the MULTIPLIER.*

T11 *connect the output of the MULTIPLIER to the CH2-A of the SCOPE SELECTOR. Adjust the oscilloscope to display the signal conveniently on the screen.*

Since each of the previous steps has been completed successfully, then at the MULTIPLIER output will be the 100% modulated AM signal. It will be displayed on CH2-A. It will look like Figure 1.

Notice the systematic manner in which the required outcome was achieved. Failure to achieve the last step could only indicate a faulty MULTIPLIER ?

agreement with theory

It is now possible to check some theory.

T12 *measure the peak-to-peak amplitude of the AM signal, with m = 1, and confirm that this magnitude is as predicted, knowing the signal levels into the MULTIPLIER, and its 'k' factor.*

the significance of 'm'

First note that the *shape* of the outline, or envelope, of the AM waveform (lower trace), is exactly that of the message waveform (upper trace). As mentioned earlier, the message includes a DC component, although this is often ignored or forgotten when making these comparisons.

You can shift the upper trace down so that it matches the envelope of the AM signal on the other trace [20]. Now examine the effect of varying the magnitude of the parameter 'm'. This is done by varying the message amplitude with the ADDER gain control **G** [21].

- for all values of 'm' less than that already set (m = 1), the envelope of the AM is the same shape as that of the message.
- for values of m > 1 the envelope is NOT a copy of the message shape.

It is important to note that, for the condition m > 1:

- it should not be considered that there is *envelope distortion*, since the resulting shape, whilst not that of the message, is the shape the theory predicts.
- there need be no *AM signal distortion* for this method of generation. Distortion of the AM signal itself, if present, will be due to amplitude overload of the hardware. But overload should not occur, with the levels previously recommended, for moderate values of m > 1.

---

[20] comparing phases is not always as simple as it sounds. With a more complex model the additional small phase shifts within and between modules may be sufficient to introduce a noticeable off-set (left or right) between the two displays. This can be corrected with a PHASE SHIFTER, if necessary.

[21] it is possible to vary the depth of modulation with either of the ADDER gain controls. But depth of modulation 'm' is considered to be proportional to the amplitude of the AC component of m(t).

T13 *vary the ADDER gain **G**, and thus 'm', and confirm that the envelope of the AM behaves as expected, including for values of m > 1.*

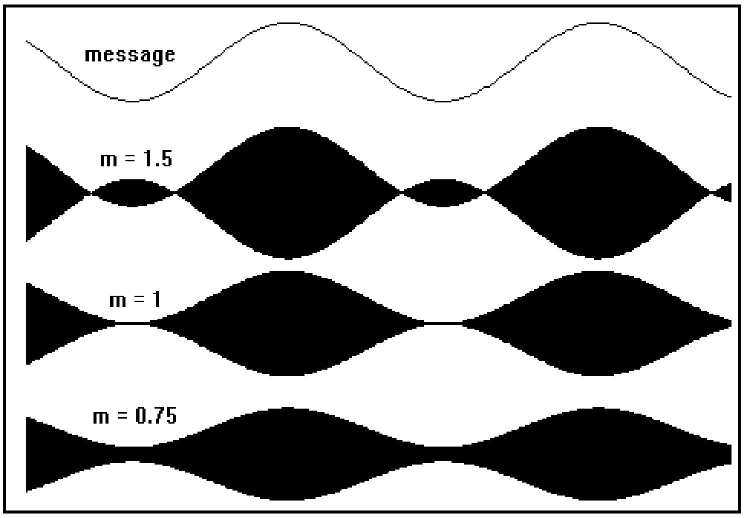


Figure 7: the AM envelope for m < 1 and m > 1

T14 *replace the AUDIO OSCILLATOR output with a speech signal available at the TRUNKS PANEL. How easy is it to set the ADDER gain **G** to occasionally reach, but never exceed, 100% amplitude modulation ?*

the modulation trapezoid

With the display method already examined, and with a sinusoidal message, it is easy to set the depth of modulation to any value of 'm'. This method is less convenient for other messages, especially speech.

The so-called *trapezoidal display* is a useful alternative for more complex messages. The patching arrangement for obtaining this type of display is illustrated in Figure 8 below, and will now be examined.

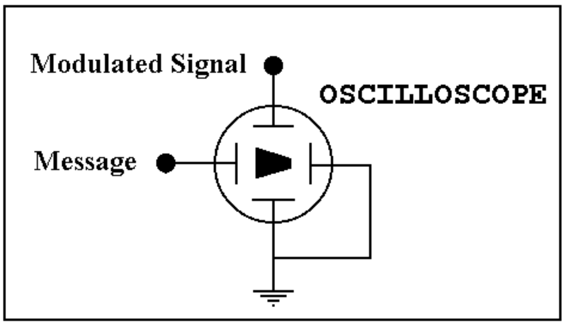


Figure 8: the arrangement for producing the TRAPEZOID

T15 *patch up the arrangement of Figure 8. Note that the oscilloscope will have to be switched to the 'X - Y' mode; the internal sweep circuits are not required.*

T16 *with a sine wave message show that, as m is increased from zero, the display takes on the shape of a TRAPEZOID (Figure 9).*

T17 *show that, for m = 1, the TRAPEZOID degenerates into a TRIANGLE*

T18 *show that, for m > 1, the TRAPEZOID extends beyond the TRIANGLE, into the dotted region as illustrated in Figure 9*

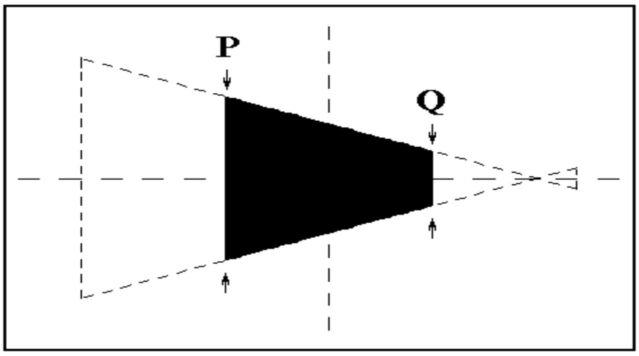


Figure 9: the AM trapezoid for m = .5. The trapezoid extends into the dotted section as m is increased to 1.2 (120%).

So here is another way of setting m = 1. But this was for a sinewave message, where you already have a reliable method. The advantage of the trapezoid technique is that it is especially useful when the message is other than a sine wave - say speech.

T19 *use speech as the message, and show that this also generates a TRAPEZOID, and that setting the message amplitude so that the depth of modulation reaches unity on peaks (a TRIANGLE) is especially easy to do.*

practical note: if the outline of the trapezoid is not made up of straight-line sections then this is a good indicator of some form of distortion. For m < 1 it could be phase distortion, but for m > 1 it could also be overload distortion. Phase distortion is not likely with TIMS, but in practice it can be caused by (electrically) long leads to the oscilloscope, especially at higher carrier frequencies.

Tutorial Questions

Q1 *there is no difficulty in relating the formula of eqn. (5) to the waveforms of Figure 7 for values of 'm' less than unity. But the formula is also valid for m > 1, provided the magnitudes P and Q are interpreted correctly. By varying 'm', and watching the waveform, can you see how P and Q are defined for m > 1 ?*

Q2 *explain how the arrangement of Figure 8 generates the TRAPEZOID of Figure 9, and the TRIANGLE as a special case.*

Q3 *derive eqn.(5), which relates the magnitude of the parameter 'm' to the peak-to-peak and trough-to-trough amplitudes of the AM signal.*

Q4 *if the AC/DC switch on the MULTIPLIER front panel is switched to AC what will the output of the model of Figure 6 become ?*

Q5 *an AM signal, depth of modulation 100% from a single tone message, has a peak-to-peak amplitude of 4 volts. What would an RMS voltmeter read if connected to this signal ? You can check your answer if you have a WIDEBAND TRUE RMS METER module.*

Q6 *in Task T6, when modelling AM, what difference would there have been to the AM from the MULTIPLIER if the opposite polarity (+ve) had been taken from the VARIABLE DC module ?*

4. ENVELOPE RECOVERY

ACHIEVEMENTS: *The ideal 'envelope detector' is defined, and then modelled. It is shown to perform well in all cases examined. The limitations of the 'diode detector', an approximation to the ideal, are examined. Introduction to the HEADPHONE AMPLIFIER module.*

PREREQUISITES: *completion of the experiment entitled* ***Envelopes*** *in this Volume.*

Preparation

the envelope

You have been introduced to the definition of an envelope in the experiment entitled *Envelopes*. There you were reminded that the envelope of a signal y(t) is that boundary within which the signal is contained, when viewed in the time domain. *It is an imaginary line.*

Although the envelope is imaginary in the sense described above, it is possible to generate, from y(t), a signal e(t), having the same shape as this imaginary line. The circuit which does this is commonly called an *envelope detector*. A better word for envelope detector would be *envelope generator*, since that is what these circuits do.

It is the purpose of this experiment for you to model circuits which will generate these envelope signals.

the diode detector

The ubiquitous *diode detector* is the prime example of an envelope generator. It is well documented in most textbooks on analog modulation. It is synonymous with the term 'envelope demodulator' in this context.

But remember: the diode detector is an *approximation* to the ideal. We will first examine the ideal circuit.

the ideal envelope detector.

The ideal envelope detector is a circuit which takes the *absolute value* of its input, and then passes the result through a *lowpass filter*. The output from this lowpass filter is the required envelope signal. See Figure 1.

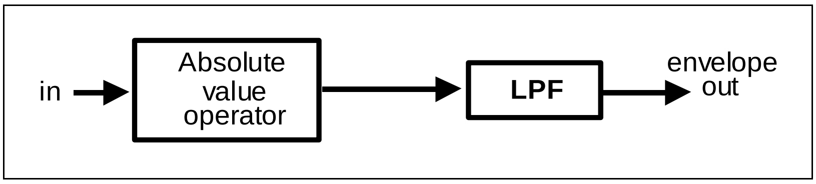


Figure 1: the ideal envelope recovery arrangement

The truth of the above statement will be tested for some extreme cases in the work to follow; you can then make your own conclusions as to its veracity.

The absolute value operation, being non-linear, must generate some new frequency components. Among them are those of the wanted envelope. Presumably, since the arrangement *actually works*, the *unwanted* components lie above those *wanted* components of the envelope.

> *It is the purpose of the lowpass filter to separate the wanted from the unwanted components generated by the absolute value operation.*

The analysis of the ideal envelope recovery circuit, for the case of a general input signal, is not a trivial mathematical exercise, the operation being non-linear. So it is not easy to define beforehand where the unwanted components lie. See the Appendix to this experiment for the analysis of a special case.

the ideal rectifier

A circuit which takes an absolute value is a fullwave rectifier. Note carefully that the operation of rectification is *non-linear*. The so-called *ideal rectifier* is a precision realization of a rectifier, using an operational amplifier and a diode in a negative feedback arrangement. It is described in text books dealing with the applications of operational amplifiers to analog circuits. An extension of the principle produces an ideal fullwave rectifier.

You will find a halfwave rectifier is generally adequate for use in an envelope recovery circuit. Refer to the Appendix to this experiment for details.

envelope bandwidth

You know what a *lowpass filter* is, but what should be its cut-off frequency in this application ? The answer: 'the cut-off frequency of the lowpass filter should be high enough to pass all the wanted frequencies in the envelope, but no more'. So you need to know the envelope bandwidth.

In a particular case you can determine the expression for the envelope from the definition given in the experiment entitled *Envelopes*, and the bandwidth by Fourier series analysis. Alternatively, you can *estimate* the bandwidth, by inspecting its shape on an oscilloscope, and then applying rules of thumb which give quick approximations.

An envelope will *always* include a constant, or DC, term.

This is inevitable from the definition of an envelope - which includes the operation of taking the absolute value. It is inevitable also in the output of a practical circuit, by the very nature of rectification.

The presence of this DC term is often forgotten. For the case of an AM signal, modulated with music, the DC term is of little interest to the listener. But it is a direct measure of the strength of the carrier term, and so is used as an automatic gain control signal in receivers.

It is *important to note* that it is possible for the bandwidth of the envelope to be much wider than that of the signal of which it is the envelope. In fact, except for the special case of the envelope modulated signal, this is generally so. An obvious example is that of the DSBSC signal derived from a single tone message.

DSBSC envelope

The bandwidth of a DSBSC signal is twice that of the highest modulating frequency. So, for a single tone message of 1 kHz, the DSBSC bandwidth is 2 kHz. But the bandwidth of the *envelope* is many times this.

For example, we know that, analytically:

$$\text{DSBSC} = \cos\mu t.\cos\omega t \qquad \text{........ } 28$$

$$= a(t).\cos[\omega_o t + \varphi(t)] \qquad \text{........ } 29$$

$$\text{because } \mu << \omega \text{ then } a(t) = \cos\mu t \qquad \text{........ } 30$$

$$\varphi(t) = 0 \qquad \text{........ } 31$$

$$\text{and envelope } e(t) = |\ a(t)\ | \quad \text{(by definition)} \qquad \text{........ } 32$$

So:

- from the mathematical definition the envelope shape is that of the absolute value of cosμt. This has the shape of a fullwave rectified version of cosμt.
- by looking at it, and from considerations of Fourier series analysis [22], the envelope must have a wide bandwidth, due to the sharp discontinuities in its shape. So the lowpass filter will need to have a bandwidth wide enough to pass at least the first few odd harmonics of the 1 kHz message; say a passband extending to *at least* 10 kHz ?

Experiment

the ideal model

The TIMS model of the ideal envelope detector is shown in block diagram form in Figure 2.

[22] see the section on *Fourier series and bandwidth estimation* in the chapter entitled ***Introduction to modelling with TIMS****, in this Volume*

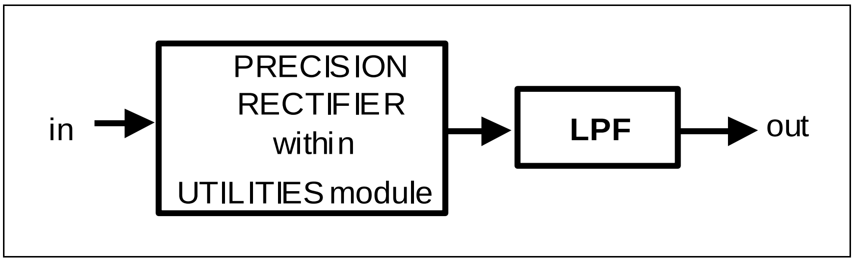


Figure 2: modelling the ideal envelope detector with TIMS

The ‘ideal rectifier’ is easy to build, does in fact approach the ideal for our purposes, and one is available as the RECTIFIER in the TIMS UTILITIES module. For purposes of comparison, a diode detector, in the form of ‘DIODE + LPF’, is also available in the same module; this will be examined later.

The desirable characteristics of the lowpass filter will depend upon the frequency components in the envelope of the signal as already discussed.

We can easily check the performance of the ideal envelope detector in the laboratory, by testing it on a variety of signals.

The actual envelope shape of each signal can be displayed by observing the modulated signal itself with the oscilloscope, suitably triggered.

The output of the envelope detector can be displayed, for comparison, on the other channel.

AM envelope

For this part of the experiment we will use the generator of Figure 3, and connect its output to the envelope detector of Figure 2.

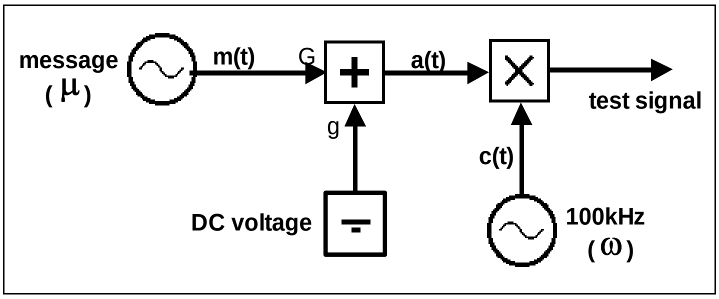


Figure 3: generator for AM and DSBSC

T1 *plug in the TUNEABLE LPF module. Set it to its widest bandwidth, which is about 12 kHz (front panel toggle switch to* WIDE*, and* TUNE *control fully clockwise). Adjust its passband gain to about unity. To do this you can use a test signal from the AUDIO OSCILLATOR, or perhaps the 2 kHz message from the MASTER SIGNALS module.*

T2 *model the generator of Figure 3, and connect its output to an ideal envelope detector, modelled as per Figure 2. For the lowpass filter use the TUNEABLE LPF module. Your whole system might look like that shown modelled in Figure 4 below.*

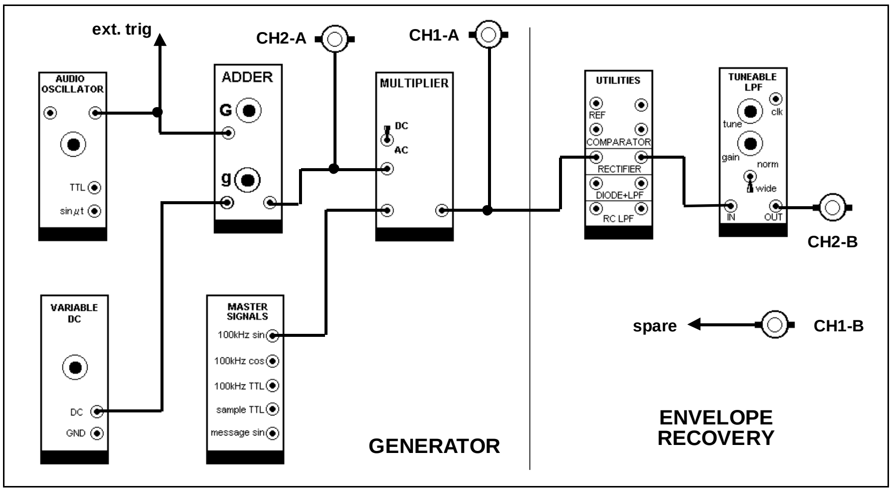


Figure 4: modulated signal generator and envelope recovery

T3 *set the frequency of the AUDIO OSCILLATOR to about 1 kHz. This is your message.*

T4 *adjust the triggering and sweep speed of the oscilloscope to display two periods of the message (CH2-A).*

T5 *adjust the generator to produce an AM signal, with a depth of modulation less than 100%. Don`t forget to so adjust the ADDER gains that its output (DC + AC) will not overload the MULTIPLIER; that is, keep the MULTIPLIER input within the bounds of the TIMS ANALOG REFERENCE LEVEL (4 volt peak-to-peak). This signal is not symmetrical about zero volts; neither excursion should exceed the 2 volt peak level.*

T6 *for the case m < 1 observe that the output from the filter (the ideal envelope detector output) is the same shape as the envelope of the AM signal - a sine wave.*

DSBSC envelope

Now let us test the ideal envelope detector on a more complex envelope - that of a DSBSC signal.

T7 *remove the carrier from the AM signal, by turning '**g**' fully anti-clockwise, thus generating DSBSC. Alternatively, and to save the DC level just used, pull out the patch cord from the '**g**' input of the ADDER (or switch the MULTIPLIER to AC).*

Were you expecting to see the waveforms of Figure 5 ? What did you see ?

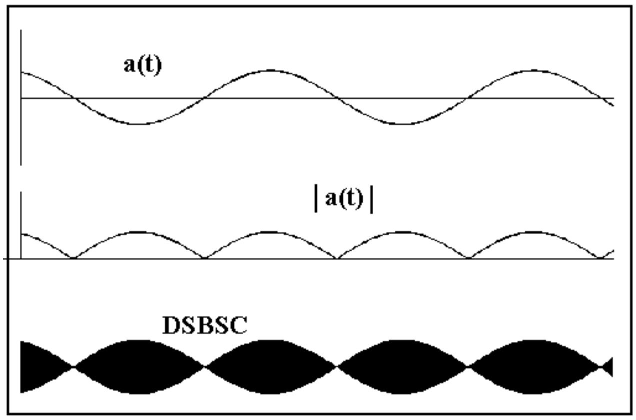


Figure 5: a DSBSC signal

You may not have seen the expected waveform. Why not ?

With a message frequency of 2 kHz, a filter bandwidth of about 12 kHz is not wide enough.

You can check this assertion; for example:

a) lower the message frequency, and note that the recovered envelope shape approaches more closely the expected shape.
b) change the filter. Try a 60 kHz LOWPASS FILTER.

T8 *(a) lower the frequency of the AUDIO OSCILLATOR, and watch the shape of the recovered envelope. When you think it is a better approximation to expectations, note the message frequency, and the filter bandwidth, and compare with predictions of the bandwidth of a fullwave rectified sinewave.*

(b) if you want to stay with the 2 kHz message then replace the TUNEABLE LPF with a 60 kHz LOWPASS FILTER. Now the detector output should be a good copy of the envelope.

speech as the message; m < 1

Now try an AM signal, with speech from a SPEECH module, as the message.

To listen to the recovered speech, use the HEADPHONE AMPLIFIER.

The HEADPHONE AMPLIFIER enables you to listen to an audio signal connected to its input. This may have come via an external lowpass filter, or via the internal 3 kHz LOWPASS FILTER. The latter is switched in and out by the front panel switch. Refer to the *TIMS User Manual* for more information.

Only for the case of envelope modulation, with the depth of modulation 100% or less, will the speech be intelligible. If you are using a separate lowpass filter, switching in the 3 kHz LPF of the HEADPHONE AMPLIFIER as well should make no difference to the quality of the speech as heard in the HEADPHONES, because the speech at TRUNKS has already been bandlimited to 3 kHz.

speech as the message; m > 1

Don't forget to listen to the recovered envelope when the depth of modulation is increased beyond 100%. This will be a distorted version of the speech.

Distortion is usually thought of as having been caused by some circuit imperfection.

> *There is* ***no circuit imperfection*** *occurring here !*

The envelope shape, for all values of m, including m > 1, is as exactly as theory predicts, using ideal circuitry.

The envelope recovery circuit you are using is close to ideal; this may not be obvious when listening to speech, but was confirmed earlier when recovering the wide-band envelope of a DSBSC.

The distortion of the speech arises quite naturally from the fact that there is a non-linear relationship between the message and the envelope, attributed directly to the absolute sign in eqn. (5).

the diode detector

It is assumed you will have referred to a text book on the subject of the *diode detector*. This is an approximation to the ideal rectifier and lowpass filter.

How does it perform on these signals and their envelopes ?

There is a DIODE DETECTOR in the UTILITIES MODULE. The diode has not been linearized by an active feedback circuit, and the lowpass filter is approximated by an RC network. Your textbook should tell you that this is a good engineering compromise in practice, provided:

a) the depth of modulation does not approach 100%
b) the ratio of carrier to message frequency is 'large'.

You can test these conditions with TIMS. The patching arrangement is simple.

T9 *connect the signal, whose envelope you wish to recover, directly to the* ANALOG INPUT *of the 'DIODE + LPF' in the UTILITIES MODULE, and the envelope (or its approximation) can be examined at the* ANALOG OUTPUT. *You should not add any additional lowpass filtering, as the true 'diode detector' uses only a single RC network for this purpose, which is already included.*

The extreme cases you could try would include:

a) an AM signal with depth of modulation say 50%, and a message of 500 Hz. What happens when the message frequency is raised ? Is $\omega >> \mu$?

b) a DSBSC. Here the inequality $\omega >> \mu$ is meaningless. This inequality applies to the case of AM with m < 1. It would be better expressed, in the present instance, as ‘he carrier frequency ω must be very much higher than the highest frequency component expected in the envelope’. This is certainly NOT so here.

T10 *repeat the previous Task, but with the RECTIFIER followed by a simple RC filter. This compromise arrangement will show up the shortcomings of the RC filter. There is an independent RC LPF in the UTILITIES MODULE. Check the* ***TIMS User Manual*** *regarding the time constant*.

T11 *you can examine various combinations of diode, ideal rectifier, RC and other lowpass filters, and lower carrier frequencies (use the VCO). The 60 kHz LPF is a very useful filter for envelope work.*

T12 *check by observation: is the RECTIFIER in the UTILITIES MODULE a halfwave or fullwave rectifier ?*

Tutorial Questions

Q1 *an analysis of the ideal envelope detector is given in the Appendix to this experiment. What are the conditions for there to be no distortion components in the recovered envelope ?*

Q2 *analyse the performance of a square-law device as an envelope detector, assuming an ideal filter may be used. Are there any distortion components in the recovered envelope ?*

Q3 *explain the major difference differences in performance between envelope detectors with half and fullwave rectifiers.*

Q4 *define what is meant by 'selective fading'. If an amplitude modulated signal is undergoing selective fading, how would this affect the performance of an envelope detector as a demodulator ?*

Appendix A

analysis of the ideal detector

The aim of the rectifier is to take the absolute value of the signal being rectified. That is, to multiply it by +1 when it is positive, and -1 when negative.

An analysis of the ideal envelope detector is not a trivial exercise, except in special cases. Such a special case is when the input signal is an envelope modulated signal with $m < 1$.

In this case we can make the following assumption, not proved here, but verified by practical measurement and observations, namely: the zero crossings of an AM signal, for $m < 1$, are uniform, and spaced at half the period of the carrier.

If this is the case, then the action of an ideal rectifier on such a signal is equivalent to multiplying it by a square wave s(t) as per Figure 1A. It is important to ensure that the phases of the AM and s(t) are matched correctly in the analysis; in the practical circuit this is done automatically.

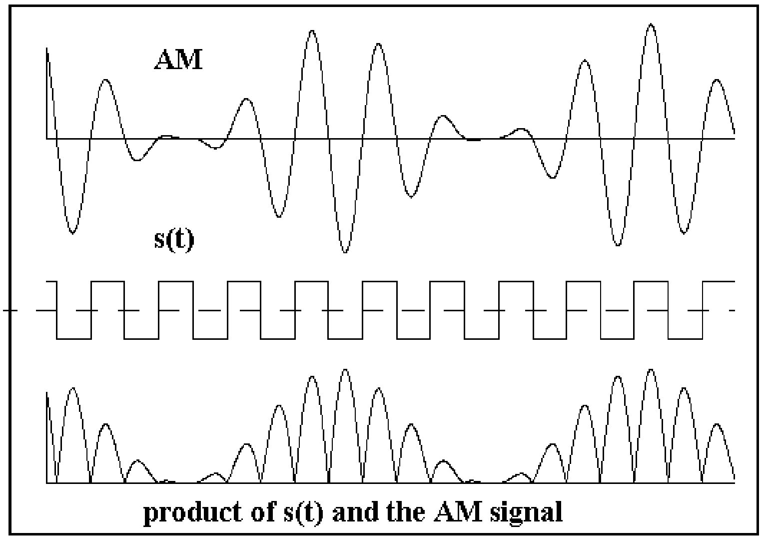


Figure 1A: the function s(t) and its operation upon an AM signal

The Fourier series expansion of s(t), as illustrated, is given by:

$$s(t) = 4/\pi\ [1.\cos\omega t - 1/3.\cos 3\omega t + 1/5.\cos 5\omega t - .....\ ] \qquad \text{.................... A1}$$

Thus s(t) contains terms in all odd harmonics of the carrier frequency

The input to the lowpass filter will be the rectifier output, which is:

$$\text{rectifier output} = s(t) \cdot \text{AM} \qquad \text{.................... A2}$$

Note that the AM is centred on ‘ω’, and s(t) is a string of terms on the ODD harmonics of ω. Remembering also that the product of two sinewaves gives ‘sum and difference’ terms, then we conclude that:

- the 1st harmonic in s(t) gives a term near DC and another centred at 2ω
- the 3rd harmonic in s(t) gives a term at 2ω and 4ω

- the 5th harmonic in s(t) gives a term at 4ω and 6ω
- and so on

We define the AM signal as:

$$AM = A\,[1 + m(t)]\cos\omega t \qquad \text{A3}$$

where, for the depth of modulation to be less than 100%, $|m(t)| < 1$.

From the rectified output we are only interested in any term near DC; this is the one we can hear. In more detail:

$$\text{term near DC} = (1/2).(4/\pi).A.m(t) \qquad \text{A4}$$

which is an *exact*, although scaled, copy of the message m(t).

The other terms are copies of the original AM, but on all even multiples of the carrier, and of decreasing amplitudes. They are easily removed with a lowpass filter. The *nearest* unwanted term is a scaled version of the original AM on a carrier frequency 2ω rad/s.

For the case where the carrier frequency is very much higher than the highest message frequency, that is when ω >> μ, an inequality which is generally satisfied, the lowpass filter can be fairly simple. Should the carrier frequency not satisfy this inequality, we can still see that the message will be recovered UNDISTORTED so long as the carrier frequency is at least twice the highest message frequency, and a filter with a steeper transition band is used.

practical modification

In practice it is easier to make a halfwave than a fullwave rectifier. This means that the expression for s(t) will contain a DC term, and the magnitudes of the other terms will be halved. The effect of this DC term in s(t) is to create an extra term in the output, namely a scaled copy of the input signal.

This is an extra unwanted term, centred on ω rad/s, and in fact the lowest frequency unwanted term. The lowest frequency unwanted term in the fullwave rectified output is centred on 2ω rad/s.

This has put an extra demand upon the lowpass filter. This is not significant when ω >> μ, but will become so for lower carrier frequencies.

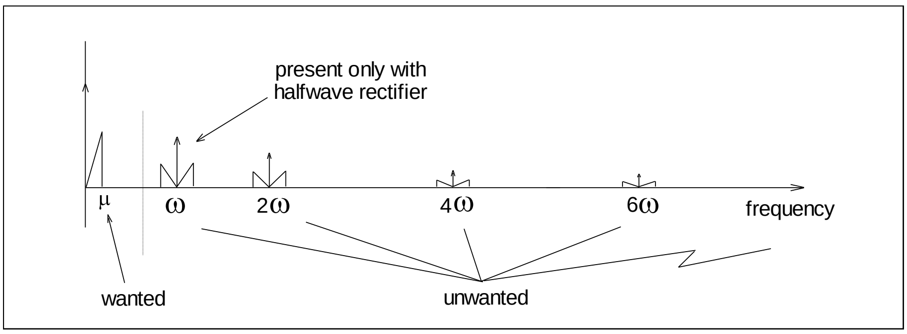


Figure 2A: rectifier output spectrum (approximate scale)

5. INTRODUCTION TO FM USING A VCO

ACHIEVEMENTS: *FM generation using a VCO. Confirmation of selected aspects of the FM spectrum. Demodulation using a zero crossing counter demodulator.*

PREREQUISITES: *familiarity with the contents of the chapter entitled* ***Analysis of the FM spectrum****; completion of the experiment entitled* ***Spectrum analysis - the WAVE ANALYSER****, both in this Volume.*

EXTRA MODULES: *SPECTRUM UTILITIES. A second VCO is required for spectral measurements.*

Introduction

This experiment has been written to satisfy those who are familiar with the existence of the ubiquitous VCO integrated circuit, and wish to explore it as a source of FM.

The VCO - *voltage controlled oscillator* - is available as a low-cost integrated circuit (IC), and its performance is remarkable. The VCO IC is generally based on a bi-stable 'flip-flop', or 'multi-vibrator' type of circuit. Thus its output waveform is a rectangular wave. However, ICs are available with this converted to a sinusoid. The mean frequency of these oscillators is determined by an RC circuit.

The *controllable* part of the VCO is its frequency, which may be varied about a mean by an external control voltage.

The variation of frequency is remarkably linear, with respect to the control voltage, over a large percentage range of the mean frequency. This then suggests that it would be ideal as an FM generator for communications purposes.

Unfortunately such is not the case.

The relative instability of the centre frequency of these VCOs renders them unacceptable for modern day communication purposes. The uncertainty of the centre frequency does not give rise to problems at the receiver, which may be taught to track the drifting carrier (see this Volume for the experiment entitled *Carrier acquisition and the PLL*, which introduces the phase locked loop - PLL). The problem is that spectrum regulatory authorities insist, and with good reason, that communication transmitters maintain their (mean) carrier frequencies within close limits [23].

It is possible to stabilise the frequency of an oscillator, relative to some fixed reference, with automatic frequency control circuitry. But in the case of a VCO which is being frequency modulated there is a conflict, with the result that the control circuitry is complex, and consequently expensive.

For applications where close frequency control is not mandatory, the VCO is used to good effect [24].

---

23 typically within a few parts per million (ppm) or less, or a few Hertz, which ever is the smaller.

24 suggest such an application

This experiment is an introduction to the FM signal. The wideband FM signal is very convenient for studying some of the properties of the FM spectrum.

spectrum

Examination of the spectrum will be carried out by modelling a WAVE ANALYSER. This instrumentation was introduced in the experiment entitled *Spectrum analysis - the WAVE ANALYSER* (in this Volume).

Specifically, two properties of an FM signal will be examined:

first Bessel zero

Check your Bessel tables and confirm that $J_0(\beta) = 0$ when $\beta = 2.45$

This Bessel coefficient controls the amplitude of the spectral component at carrier frequency, so with $\beta = 2.45$ there should be a carrier null.

You will be able to set $\beta = 2.45$ and so check this result.

special case - β = 1.45

For $\beta = 1.45$ the amplitude of the first pair of sidebands is equal to that of the carrier; and this will be $J_0(1.45)$ times the amplitude of the unmodulated carrier (always available as a reference).

You should confirm this result from Tables of Bessel functions (see, for example, the Appendix C to this volume).

This is one of the many special cases one can examine, to further verify the predictions of theory.

the zero-crossing-counter demodulator

A simple yet effective FM demodulator is one which takes a time average of the zero crossings of the FM signal. Figure 1 suggests the principle.

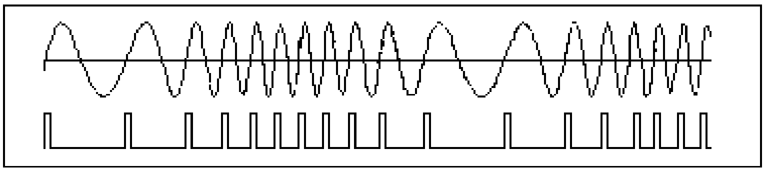


Figure 1: an FM signal, and a train of zero-crossing pulses

Each pulse in the pulse train is of *fixed width*, and is located at a zero crossing of the FM signal [25]. This is a pulse-repetition-rate modulated signal. If the pulse train is passed through a lowpass filter, the filter will perform an averaging operation. The rate of change of this average value is related to the message frequency, and the magnitude of the change to the depth of modulation at the generator.

[25] only positive going zero crossings have been selected

This zero-crossing-counter demodulator [26] will be modelled in the latter part of the experiment. The phase locked loop (PLL) as a demodulator is studied in the experiment entitled *FM demodulation and the PLL* in this Volume.

Experiment

A suitable set-up for measuring some properties of a VCO is illustrated in Figure 2.

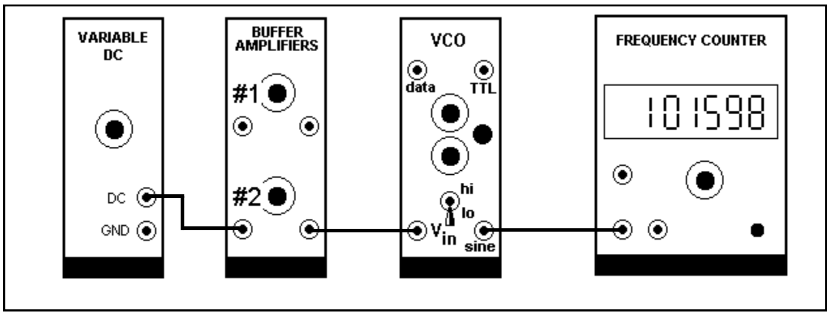


Figure 2: the FM generator

For this experiment you will need to measure the sensitivity of the frequency of the VCO to an external control voltage, so that the frequency deviation can be set as desired.

The mean frequency of the VCO is set with the front panel control labelled f_o.

The mean frequency can as well be varied by a DC control voltage connected to the V_{in} socket. Internally this control voltage can be amplified by an amount determined by the setting of the front panel GAIN control. Thus the frequency sensitivity to the external control voltage is determined by the GAIN setting of the VCO.

A convenient way to set the sensitivity (and thus the GAIN control, which is not calibrated), to a definable value, is described below.

T1 *before plugging in the VCO, set the mode of operation to 'VCO' with the on-board switch SW2. Set the front panel switch to 'LO'. Set the front panel* GAIN *control fully anti-clockwise.*

T2 *patch up the model of Figure 2.*

T3 *use the FREQUENCY COUNTER to monitor the VCO frequency. Use the front panel control f_o to set the frequency to 10 kHz.*

deviation sensitivity

T4 *set the VARIABLE DC module output to <u>about</u> +2 volt. Connect this DC voltage, via* BUFFER *#2, to the V_{in} socket of the VCO.*

[26] also called the *zero crosssing detector*

T5 *with the BUFFER #2 gain control, set the DC at the VCO V_{in} socket to <u>exactly</u> -1.0 volt. With the VCO GAIN set fully anti-clockwise, this will have no effect on f_0.*

T6 *increase the VCO GAIN control from zero until the frequency changes by 1 kHz. Note that the direction of change will depend upon the polarity of the DC voltage.*

> *The GAIN control of the VCO **is now set** to give a 1 kHz peak frequency deviation for a modulating signal at V_{in} of 1 volt peak.*

The gain control setting will now remain unchanged.

For this setting you have calibrated the sensitivity, ***S***, of the VCO for the purposes of the work to follow.

Here:

$$\boldsymbol{S} = 1000 \text{ Hz/volt} \qquad \text{........ } 33$$

deviation linearity

The linearity of the modulation characteristic can be measured by continuing the above measurement over a range of input DC voltages. If a curve is plotted of DC volts versus frequency deviation the linear region can be easily identified.

A second, dynamic, method would be to use a demodulator, using an audio frequency message. This will be done later. In the meantime:

T7 *take a range of readings of frequency versus DC voltage at V_{in} of the VCO, sufficient to reveal the onset of non-linearity of the characteristic. This is best done by producing a plot as the readings are taken.*

the FM spectrum

So far you have a *theoretical* knowledge of the spectrum of the signal from the VCO, but have made no *measurement* to confirm this.

The instrumentation to be modelled is the WAVE ANALYSER, introduced in the experiment entitled *Spectrum analysis - the WAVE ANALYSER*. It is assumed you have completed that experiment.

the WAVE ANALYSER

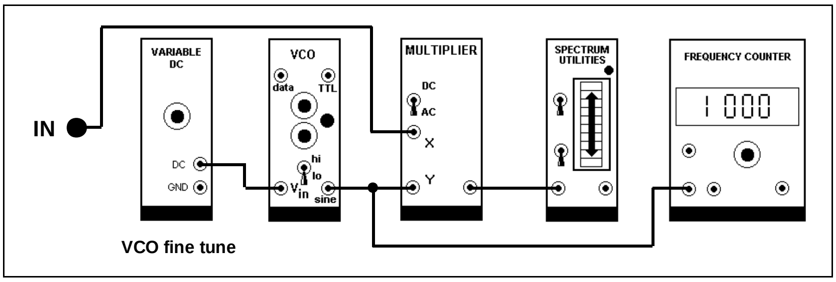


Figure 3: the WAVE ANALYSER model.

The VARIABLE DC voltage, with a moderate setting of the VCO GAIN control, is used as a fine tuning control.

Remember that one is generally not interested in absolute amplitudes - what is sought are *relative* amplitudes of spectral components.

Two such spectra will now be studied.

1. the first Bessel zero of the carrier term will be set up.
2. the amplitude of the carrier will be made equal to that of each of the first pair of sidebands.

The VCO will be set up with a sinusoidal message of 500 Hz.

T8 *in the FM generator model of Figure 2 replace the variable DC module with an AUDIO OSCILLATOR, tuned to 500 Hz.*

T9 *patch up the WAVE ANALYSER of Figure 3.*

T10 *revise your skills with the WAVE ANALYSER. Set the frequency deviation of the FM VCO to zero, with the BUFFER amplifier #2, and search for the carrier component. Set the on-board SCALING resistor of the SPECTRUM UTILITIES module to obtain a full scale deflection (FSD) on the carrier.*

first Bessel zero

Refer to the work, under this heading, which was done preparatory to starting the experiment.

Since

$$\beta = \Delta f / f_m$$

and

$$\Delta f = V_{in} \boldsymbol{S}$$

then set

$$V_{in} = \beta.f_m / \boldsymbol{S} \cong 1.22 \text{ peak volt}$$

T11 *adjust, with the AUDIO OSCILLATOR supplying the message, via the BUFFER amplifier, for β = 2.45*

T12 *use the WAVE ANALYSER to confirm the amplitude of the carrier has fallen very low. Fine trim of the BUFFER amplifier gain control should find the null. Check that β is still close to 2.45.*

T13 *find one of the adjacent sidebands. Its amplitude should be $J_1(\beta)$ times the amplitude of the unmodulated carrier (measured previously).*

since $J_1(2.45) \cong 0.5$

then each of the first pair of sideband should be of amplitude half that of the unmodulated carrier.

special case - β = 1.45

Refer to the work, under this heading, which was done preparatory to starting the experiment.

T14 *set β = 1.45 and confirm that the carrier component, and either or both of the first pair of sideband, are of similar amplitude.*

T15 *check that each of the components just measured is $J_1(1.45)$ times the amplitude of the unmodulated carrier.*

There are many more special cases, of a similar nature, which you could investigate. The most obvious would be the Bessel zeros of some of the side-frequencies.

You will now use this generator to provide an input to a demodulator.

FM demodulation

A simple FM demodulator, if it reproduces the message without distortion, will provide further confirmation that the VCO output is indeed an FM signal.

A scheme for achieving this result was introduced earlier - the zero-crossing-counter demodulator - and is shown modelled in Figure 4.

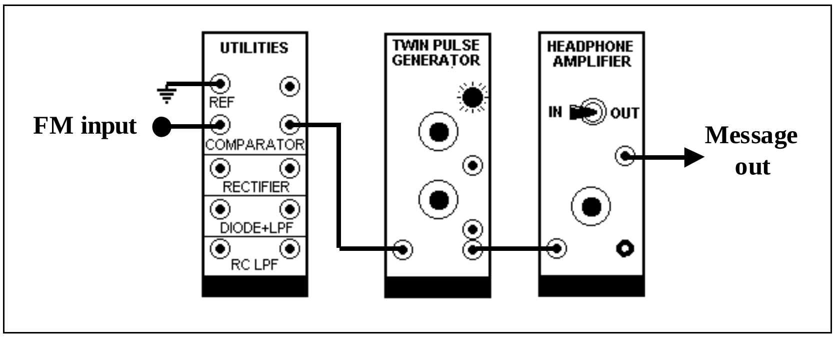


Figure 4: an FM demodulator using a zero-crossing demodulator

The TWIN PULSE GENERATOR is required to produce a pulse at each positive going zero crossing of the FM signal. To achieve this the FM signal is converted to a TTL signal by the COMPARATOR, and this drives the TWIN PULSE GENERATOR.

note: the input signal to the HEADPHONE AMPLIFIER filter is at *TTL* level. It is TIMS practice, in order to avoid overload, not to connect a TTL signal to an analog input. Check for overload. If you prefer you can use the yellow analog output from the TWIN PULSE GENERATOR. This is an AC coupled version of the TTL signal.

T16 *before plugging in the TWIN PULSE GENERATOR set the on-board MODE switch SW1 to SINGLE. Patch up the demodulator of Figure 4.*

T17 *set the frequency deviation of the FM generator to zero, and connect the VCO output to the demodulator input.*

T18 *using the WIDTH control of the TWIN PULSE GENERATOR adjust the output pulses to their maximum width.*

T19 *observe the demodulator output. If you have chosen to take the TTL output from the TWIN PULSE GENERATOR there should be a DC voltage present. Why ? Notice that it is proportional to the width of the pulses into the LPF of the HEADPHONE AMPLIFIER.*

T20 *introduce some modulation at the VCO with the BUFFER amplifier gain control. Observe the output from the LPF of the HEADPHONE AMPLIFIER using the oscilloscope. Measure its frequency (and compare with the message source at the transmitter).*

T21 *show that the amplitude of the message output from the demodulator:*

a) <u>varies</u> with the message amplitude into the VCO. Is this a linear variation ?

b) <u>varies</u> with the pulse width from the TWIN PULSE GENERATOR. Is this a linear variation ?

c) is <u>constant</u> with the frequency of the message to the VCO. Does this confirm the VCO is producing FM, and not PM ?

T22 *increase the message amplitude into the VCO until distortion is observed at the receiver output. Can you identify the source of this distortion ? Record the amplitude of the message at the VCO. You may need to increase the GAIN of the VCO.*

conclusions

There are many other observations you could make.

- have you checked (by calculation) the bandwidth of the FM signal under all conditions ? Could it ever extend ‘below DC’ and cause wrap-around (fold-back) problems ?
- do you think there is any conflict with the nearness of the message frequency to the carrier frequency ? Why not increase the carrier frequency to the limit of the VCO on the ‘LO’ range - about 15 kHz
- why not avoid possible problems caused by the relatively large ratio μ/ω and change to the 100 kHz region ?

- try a more demanding distortion test with a two-tone message.

Tutorial Questions

Q1 *name some applications where moderate carrier instability of an FM system is acceptable.*

Q2 *draw the amplitude/frequency spectrum of the signal generated in task **T14**.*

Q3 *how would you define the bandwidth of the signal you generated in Task **T14** ?*

Q4 *what will the FREQUENCY COUNTER indicate when connected to the FM signal from the VCO ? Discuss possibilities.*

Q5 *derive an expression for the sensitivity of the demodulator of Figure 4, and compare with measurements.*

$$sensitivity = \frac{output\ (message)\ amplitude}{input\ (FM)\ \ frequency\ \ deviation}$$

Q6 *what is a magnitude for β, in $J_n(\beta)$, for a Bessel zero, if:*

a) $n = 1$
b) $n = 3$

6. FM DEMODULATION WITH THE PLL

ACHIEVEMENTS: *introduction to the PLL as an FM demodulator*

PREREQUISITES: *an understanding of the contents of the Chapter entitled* ***Analysis of the FM spectrum*** *(this Volume) is desirable, but not essential. A familiarity with the analysis of a PLL will allow some quantitative measurements to be made and interpreted.*

preparation

the phase locked loop - PLL.

The phase locked loop is a non-linear feedback loop. To analyse its performance to any degree of accuracy is a non-trivial exercise. To illustrate it in simplified block diagram form is a simple matter. See Figure 1.

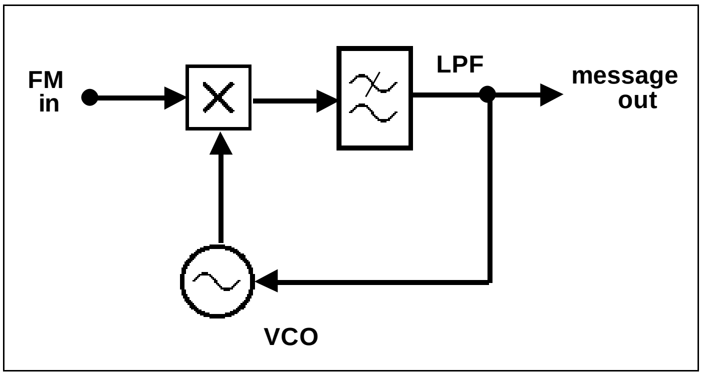


Figure 1: the basic PLL

This arrangement has been used in an earlier experiment (this Volume), namely that entitled *Carrier acquisition and the PLL*, where an output was taken from the VCO. As an FM demodulator, the output is taken from the LPF, as shown.

It is a simple matter to describe the principle of operation of the PLL as a demodulator, but another matter to carry out a detailed analysis of its performance. It is complicated by the fact that its performance is described by non-linear equations, the solution to which is generally a matter of approximation and compromise.

The principle of operation is simple - or so it would appear. Consider the arrangement of Figure 1 in open loop form. That is, the connection between the filter output and VCO control voltage input is broken.

Suppose there is an unmodulated carrier at the input.

The arrangement is reminiscent of a product, or multiplier-type, demodulator. If the VCO was tuned precisely to the frequency of the incoming carrier, ω_0 say, then the output would be a DC voltage, of magnitude depending on the phase difference between itself and the incoming carrier.

For two angles within the 360^0 range the output would be precisely zero volts DC.

Now suppose the VCO started to drift slowly off in frequency. Depending upon which way it drifted, the output voltage would be a slowly varying AC, which if slow enough looks like a varying amplitude DC. The sign of this DC voltage would depend upon the direction of drift.

Suppose now that the loop of Figure 1 is closed. If the sign of the slowly varying DC voltage, now a VCO *control voltage*, is so arranged that it is in the direction to urge the VCO back to the incoming carrier frequency ω_0, then the VCO would be encouraged to 'lock on' to the incoming carrier.

This is the principle of carrier acquisition. This was examined in the experiment entitled *Carrier acquisition and the PLL*, where this same description was used.

Next suppose that the incoming carrier is frequency modulated. For a low frequency message, and small deviation, you can imagine that the VCO will endeavour to follow the incoming carrier frequency. What about wideband FM ? With 'appropriate design' of the lowpass filter and VCO circuitry the VCO will follow the incoming carrier for this too.

The control voltage to the VCO will endeavour to keep the VCO frequency locked to the incoming carrier, and thus will be an exact copy of the original message.

Rather than attempt to analyse the operation of the arrangement of Figure 1 as a demodulator, you will make a model of it, and demonstrate that it is able to recover the message from an FM signal.

Experiment

FM demodulation

There is an FM signal at TRUNKS. It is based on a nominal 100 kHz carrier. You will model the PLL, and recover the message from the FM signal.

T1 *make a model of the PLL of Figure 1. Use the RC-LPF in the UTILITIES module. Remember to set up the VCO module in 100 kHz VCO mode. In the first instance set the front panel* GAIN *control to its mid-range position.*

T2 *examine, with your oscilloscope, the FM signal at TRUNKS. Identify those features which suggest it could indeed be an FM signal.*

T3 *connect the FM signal at TRUNKS to the PLL.*

The PLL may or may not at once lock on to the incoming FM signal. This will depend upon several factors, including:

- the frequency to which the PLL is tuned
- the capture range of the PLL
- the PLL loop gain - the setting of the front panel GAIN control of the VCO

You will also need to know what method you will use to verify that lock has taken place.

When you have satisfied yourself that you understand the significance of these considerations then you should proceed.

T4 *make any necessary adjustments to the PLL to obtain lock, and record how this was done. Measure the amplitude and frequency of the recovered message (if periodic), or otherwise describe it (speech or music ?). Are any of these measurements dependent upon the setting of the VCO* GAIN *control ?*

If you are familiar with the analysis of the PLL you should complete the next task.

T5 *measure the properties of each element of the PLL, and then predict some of its properties as a demodulator. If the message was a single tone, from its amplitude can you estimate the frequency deviation of the FM signal ?*

more measurements

If you have two VCO modules you can make your own FM signal. You will then have access to both the original message and the demodulator output. This will allow further measurements.

T6 *set up an FM signal, using a VCO, as described in the experiment entitled* ***Introduction to FM using a VCO****. Use any suitable message frequency, and a frequency deviation of say 5 kHz.*

T7 *compare the waveform and frequency of the message at the transmitter, and the message from the demodulator.*

T8 *check the relationship between the message amplitude at the transmitter, and the message amplitude from the demodulator.*

T9 *as a further confidence check, use the more demanding two-tone signal as a test message. The two tones can come from an AUDIO OSCILLATOR and the 2.033 kHz message from the MASTER SIGNALS module, combined in an ADDER.*

Tutorial Questions

Even if you are unable to complete any of the following questions in detail, you should read them, and a text book on the subject, so as to obtain some appreciation of the behaviour of the PLL, especially as an FM demodulator.

Q1 *define capture range, lock range, demodulation sensitivity of the PLL as an FM demodulator. What other parameters are important ?*

Q2 *how does the sensitivity of the VCO, to the external modulating signal, determine the performance of the demodulator ?*

Q3 *what is the significance of the bandwidth of the LPF in the phase locked loop ?*

7. SAMPLING WITH SAMPLE AND HOLD

ACHIEVEMENTS: *investigation of the sample-and-hold operation as a first step towards digitization of an analog waveform. Message reconstruction by lowpass filtering.*

PREREQUISITES: *none*

ADVANCED MODULES: *INTEGRATE & DUMP*

Preparation

A/D conversion

Before it is possible to transmit analog information via a digital system the analog signal must first be transformed into a digital format. The *first step* in such a transformation typically involves a sampling process.

natural sampling

Natural sampling of an analog waveform (message) is examined in the experiment entitled *The sampling theorem* (within *Volume A1 - Fundamental Analog Experiments*).

Natural sampling takes a slice of the waveform, and the top of the slice preserves the shape of the waveform.

flat top sampling

A very common, and easily implemented method of sampling of an analog signal uses the sample-and-hold operation. This produces *flat top* samples.

Flat top sampling takes a slice of the waveform, but cuts off the top of the slice horizontally. The top of the slice does *not* preserve the shape of the waveform.

Figure 1 below contrasts the two methods.

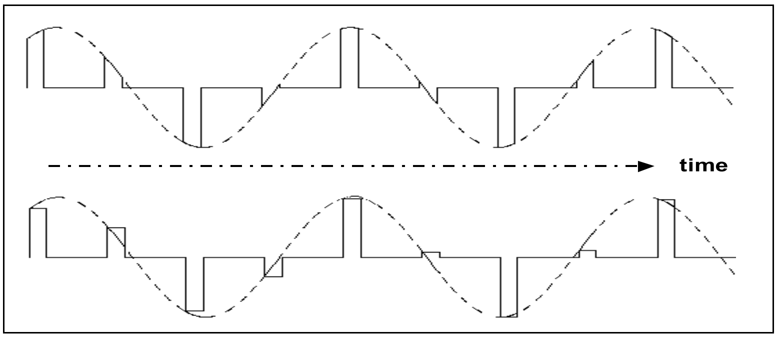


Figure 1: natural sampling (above) and flat top (below)

message reconstruction by lowpass filtering

In the experiment entitled *The sampling theorem* a simple analysis showed that there was *no distortion* of the message when reconstruction was implemented by lowpass filtering.

It will now be declared as an obvious fact:

> *if message reconstruction by lowpass filtering of natural samples results in no distortion, then there must be distortion when flat top pulses are involved.*

Analysis of the distortion for flat top pulses will not be attempted here. Instead some observations will be made, and you can draw your own conclusions.

sample width

An important observation must be made. The pulse width determines the amount of energy in each pulse, and so can determine the amplitude of the reconstructed message. But, in a linear and noise free system, the width of the samples plays no part in determining the amount of distortion of a reconstructed message.

sample-and-hold sampling

The sample-and-hold operation is simple to implement, and is a very commonly used method of sampling in communications systems.

In its simplest form the sample is held until the next sample is taken. So it is of maximum width.

This is illustrated in Figure 2 below.

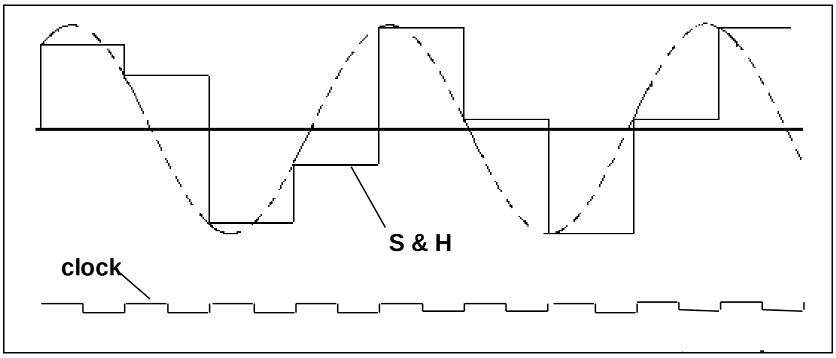


Figure 2: sampling by sample-and-hold (for full sample width)

In the above example the sampling instant is coincident with the rising edge of the clock signal.

In practice there may be a 'processing delay' before the stepped waveform is presented at the output. This is the case in the sub-system being examined in this experiment.

Experiment

There is a stand alone SAMPLE-AND-HOLD sub-system in the INTEGRATE & DUMP module. This will be used in the present experiment.

T1 *acquire an INTEGRATE & DUMP module. This is a multi-purpose module. Within it is a sub-system which performs sample-and-hold operations. Before plugging it in, set the on-board switch SW1 to the S&H 1 position ('0'). Analog signals connected to the input socket labelled I&D 1 will now undergo a sample-and-hold (S&H 1) operation, the result appearing at the I&D 1 output socket. Ignore the duplicate S&H 2 option available at the I&D 2 sockets.*

T2 *patch up the module according to Figure 3 below.*

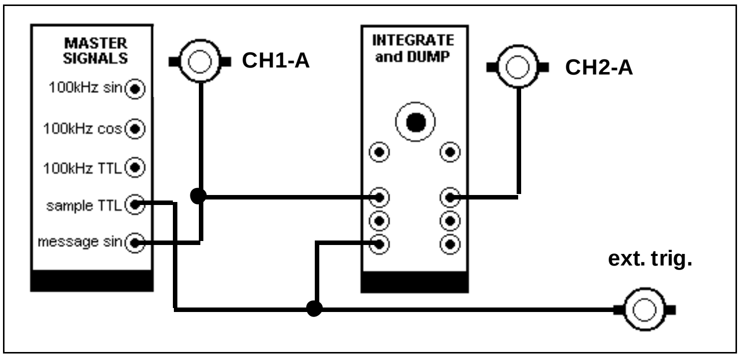


Figure 3: the TIMS model

For a stable view of both input and output it is convenient to use a message which is a submultiple of the sample clock frequency. Thus use the 2.03 kHz message (sinewave) from the MASTER SIGNALS module, together with the 8.333 kHz TTL clock.

T3 *select a sweep speed to show two or three periods of the message - say 0.1 ms/cm. Set equal gains of both channels - say 1 volt/cm. With the patching shown in Figure 3 you might expect to obtain oscilloscope displays similar to that of Figure 2. Try it.*

T4 *note the output from the socket labelled READY. Sketch it with respect to the clock and output signal, showing time relationships.*

There is a processing delay within the sample-and-hold sub-system. As a result, the two displays will be shifted relative in time. The ready signal occurs within the time during which the sample is available, and could be used to signal analog-to-digital (A/D) circuitry to start a conversion.

message reconstruction

Now that you have seen a sample-and-hold operation, you are ready to reconstruct the message from it. This is a lowpass filtering operation.

T5 *use a TUNEABLE LPF module to reconstruct the message. Decide on, then set, a 'suitable' bandwidth. Report your findings. Then read on:*

To what passband width did you set the filter ?

Remember, you are looking for any possible distortion components introduced by the sample-and-hold operation, and then the reconstruction process.

Since the message is at 2.03 kHz a passband of 3 kHz would be wide enough ?

Yes and no !

This would indeed be wide enough to pass the message, but it would not be wide enough to pass any harmonic distortion components.

But the filter passband could not be made wider than half the sampling frequency (else the Nyquist criterion would be violated), and that is not much more than the current message frequency. So something has to be changed.

Is a synchronous message necessary ? Not any longer, after having seen the stationary sample-and-hold waveform. So why not use an AUDIO OSCILLATOR, set to its lowest frequency (about 300 Hz), and the 3 kHz LPF within the HEADPHONE AMPLIFIER module. This would give plenty of room for any distortion components to appear at the output. However, unless they are of significant amplitude, they may not be visible on the oscilloscope.

T6 *do as suggested above. Use the oscilloscope to view both the input and output sinewaves simultaneously. Synchronize the oscilloscope (externally) to the source of the message. As an engineering estimate, if the distortion is not obvious, then one could say the signal-to-distortion ratio is better than 30 dB (probably better than 40 dB).*

As well as one can judge the two waveforms are 'identical' ? Could you estimate the amount of distortion introduced by the reconstruction process ? See Tutorial Question Q1.

If there *was* visible distortion then one should check the 3 kHz LPF reconstruction filter - does it introduce its own distortion ? Compare the message shape *before* sampling, *but via this filter*, as well as *after* reconstruction.

Could you attempt to *measure* the mount of distortion ? See Tutorial Question Q2. The unwanted components will probably be hidden in the noise level; meaning the signal-to-distortion ratio is much better than 40 dB.

two-tone test signal

Testing for distortion with a single sine wave is perhaps not demanding enough. Should you try a two-tone test signal ? The technique was introduced in the experiment entitled *Amplifier overload* (within *Volume A2 - Further & Advanced Analog Experiments*).

aliasing

With the 3 kHz LPF as the reconstruction filter, and an 8.333 kHz sample rate, there should be no sign of aliasing distortion.

To demonstrate aliasing distortion:

T7 *replace the 8.333 kHz sampling signal from the MASTER SIGNALS module with the TTL output from a VCO. Monitor the VCO frequency with the FREQUENCY COUNTER. Starting with the VCO set to its highest frequency on the LO range (about 15 kHz), slowly reduce it, while watching the reconstructed message waveshape. As soon as distortion is evident note the VCO frequency. Knowing the reconstruction filter amplitude characteristic, how does this agree with the Nyquist criterion ? See Tutorial Question Q3.*

conclusion

You have seen that the sample-and-hold operation followed by a lowpass filter can reconstruct the signal, whose samples were taken, with ‘good’ accuracy. If you had available a spectrum analyser, or its equivalent, you would have been able to show that unwanted components were at least 40 dB below the wanted components when implemented with TIMS modules operating within their limits. So, for communications purposes, we might say message reconstruction is distortionless.

The sampling process is the first of two major steps in preparing an analog message for digital transmission. The second step is conversion of the sampled waveform to a series of digital numbers. This introduces a second source of distortion, due to the need for *quantization*. But quantization distortion can also be made negligible if sufficient quantization levels are used.

Sample-and-hold followed by amplitude quantization is examined in the experiments entitled *PCM encoding* and *PCM decoding* in this Volume.

Tutorial Questions

Q1 *assuming a sinewave is accompanied by a small third harmonic component, how large would this have to be before its presence could be detected using only an oscilloscope ? This question would not please the purists, because it raises more questions than it asks. But attempt an answer. You could even set up the signal using TIMS and* ***demonstrate*** *your reply.*

Q2 *recall the experiment entitled* **Modelling an equation** *within Volume A1 - Fundamental Analog Experiments. There was demonstrated the cancellation of a component in a signal. Describe how this technique might be used in the present case to make a measurement of signal-to-distortion ratio.*

Q3 *define the 'slot bandwidth' of a lowpass filter. Redefine the Nyquist criterion in terms of practical filter characteristics [27].*

Q4 *sample-and-hold (flat-top sampling) can be shown to introduce distortion of the message if it is reconstructed by using a lowpass filter alone. From your general reading, or otherwise, is it possible to eliminate this distortion by further message processing ?* ***hint****: key words are aperture effect, sinx/x correction.*

[27] filter characteristics are defined in Appendix A of Volume A1.

8. PCM ENCODING

ACHIEVEMENTS: *introduction to pulse code modulation (PCM) and the PCM ENCODER module. Coding of a message into a train of digital words in binary format.*

PREREQUISITES: *an understanding of sampling, from previous experiments, and of PCM from course work or a suitable text. Completion of the experiment entitled* ***Sampling with SAMPLE & HOLD*** *(in this Volume) would be a distinct advantage.*

ADVANCED MODULES: *PCM ENCODER*

Preparation

PCM

This is an introductory experiment to pulse code modulation - PCM.

The experiment will acquaint you with the PCM ENCODER, which is one of the TIMS Advanced Modules. This module generates a PCM output signal from an analog input message.

In this experiment the module will be used in isolation; that is, it will not be part of a larger system. The formatting of a PCM signal will be examined in the time domain.

A later experiment, entitled *PCM decoding* (in this Volume), will illustrate the recovery of the analog message from the digital signal.

In another experiment, entitled *PCM TDM* (within *Volume D2 - Further & Advanced Digital Experiments*), the module will be part of a system which will generate a two-channel pulse code modulated time division multiplexed system (PCM TDM).

PCM encoding

The input to the PCM ENCODER module is an analog message. This must be constrained to a defined bandwidth and amplitude range.

The maximum allowable message bandwidth will depend upon the sampling rate to be used. The Nyquist criterion must be observed.

The amplitude range must be held within the ± 2.0 volts range of the TIMS ANALOG REFERENCE LEVEL. This is in keeping with the input amplitude limits set for all analog modules.

A step-by-step description of the operation of the module follows:

1. the module is driven by an external TTL clock.
2. the input analog message is *sampled* periodically. The *sample rate* is determined by the external clock.

3. the sampling is a *sample-and-hold* operation. It is internal to the module, and cannot be viewed by the user [28]. What is held is the *amplitude* of the analog message *at the sampling instant*.
4. each sample amplitude is compared with a finite set of amplitude levels. These are distributed (uniformly, for *linear* sampling) within the range ± 2.0 volts (the TIMS ANALOG REFERENCE LEVEL). These are the system *quantizing* levels.
5. each quantizing level is assigned a *number*, starting from zero for the lowest (most negative) level, with the highest number being (L-1), where L is the available number of levels.
6. each sample is *assigned* a digital (binary) code word representing the number associated with the quantizing level which is closest to the sample amplitude. The number of bits 'n' in the digital code word will depend upon the number of quantizing levels. In fact, $n = \log_2(L)$.
7. the code word is *assembled into a time frame* together with other bits as may be required (described below). In the TIMS PCM ENCODER (and many commercial systems) a single extra bit is added, in the least significant bit position. This is alternately a *one* or a *zero*. These bits are used by subsequent decoders for frame synchronization.
8. the *frames* are transmitted serially. They are transmitted at the same rate as the samples are taken (but see Tutorial Question 3). The serial bit stream appears at the output of the module.
9. also available from the module is a synchronizing signal FS ('frame synch'). This signals the *end* of each data frame.

[28] the *sample and hold* operation is examined separately in the experiment entitled *Sampling with SAMPLE & HOLD* in this Volume.

the PCM ENCODER module

front panel features

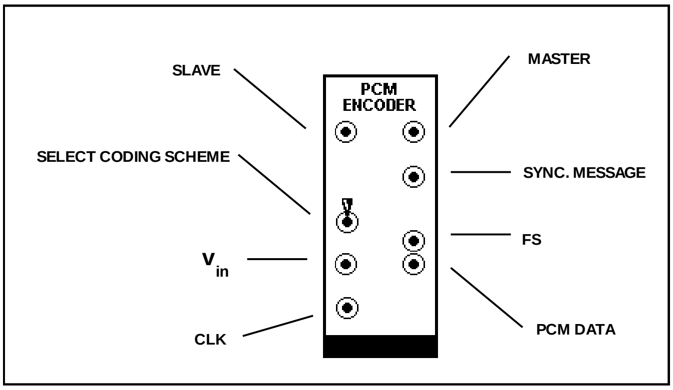


Figure 2: front panel layout of the PCM ENCODER

The front panel layout of the module is shown in Figure 2. Technical details are described in the TIMS *Advanced Modules User Manual*.

Note and understand the purpose of each of the input and output connections, and the three-position toggle switch. Counting from the top, these are:

- ***SLAVE***: not used during this experiment. Do *not* connect anything to this input.
- ***MASTER***: not used during this experiment. Do *not* connect anything to this output.
- ***SYNC. MESSAGE****:* periodic, 'synchronized', message. Either sinusoidal, or sinusoidal-like ('sinuous'), its frequency being a sub-multiple of the MASTER CLOCK (being any one of four frequencies selected by an on-board switch SW2). A message synchronized to the system clock is convenient for obtaining stable oscilloscope displays. Having a recognisable shape (but being more complex than a simple sine wave) gives a qualitative idea of distortion during the decoding process (examined in a later experiment). See Table A-1 in the Appendix to this experiment for more details.
- ***SELECT CODING SCHEME****:* a three-position toggle switch which selects the 4-bit or 7-bit encoding scheme of the analog samples; or (together with an on-board jumper connection) the companding scheme.
- ***FS***: frame synchronization, a signal which indicates the end of each data frame.
- ***V_{in}:***: the analog signal to be encoded.
- ***PCM DATA***: the output data stream, the examination of which forms the major part of this experiment.
- ***CLK***: this is a TTL (red) input, and serves as the MASTER CLOCK for the module. Clock rate must be 10 kHz or less. For this experiment you will use the 8.333 kHz TTL signal from the MASTER SIGNALS module.

the TIMS PCM time frame

Each binary word is located in a *time frame*. The time frame contains eight *slots* of equal length, and is eight clock periods long. The slots, from first to last, are numbered 7 through 0. These slots contain the bits of a binary word. The least significant bit (LSB) is contained in slot 0.

The LSB consists of alternating *ones* and *zeros*. These are placed ('embedded') in the frame by the encoder itself, and cannot be modified by the user. They are used by subsequent decoders to determine the location of each frame in the data stream, and its length. See the experiment entitled *PCM decoding* (in this Volume).

The remaining seven slots are available for the bits of the binary code word. Thus the system is capable of a resolution of seven-bits maximum. This resolution, for purposes of experiment, can be reduced to four bits (by front panel switch). The 4-bit mode uses only five of the available eight slots - one for the embedded frame synchronization bits, and the remaining four for the binary code word (in slots 4, 3, 2, and 1).

pre-calculations

You will be using an 8.333 kHz master clock. Answer Tutorial Question Q1 now, *before* commencing the experiment.

Experiment

The only module required for this experiment is a TIMS PCM ENCODER.

It is not necessary, for this experiment, to become involved with *how* the PCM ENCODER module achieves its purpose. What will be discovered is *what* it does under various conditions of operation.

The module is capable of being used in two modes: as a stand-alone PCM encoder, for one channel, or, with modifications to the data stream, as part of a two-channel time division multiplexed (TDM) PCM system.

Operation as a single channel PCM encoder is examined in this experiment.

Before plugging the module in:

T1 *select the TIMS companding A_4-law with the on-board COMP jumper (in preparation for a later part of the experiment).*

T2 *locate the on-board switch SW2. Put the LEFT HAND toggle DOWN and the RIGHT HAND toggle UP. This sets the frequency of a message from the module at SYNC. MESSAGE. This message is synchronized to a sub-multiple of the MASTER CLOCK frequency. For more detail see the Appendix to this experiment.*

patching up

To determine some of the properties of the analog to digital conversion process it is best to start with a DC message. This ensures completely stable oscilloscope displays, and enables easy identification of the quantizing levels.

Selecting the 4-bit encoding scheme reduces the number of levels (2^4) to be examined.

T3 *insert the module into the TIMS frame. Switch the front panel toggle switch to 4-BIT LINEAR (ie., no companding).*

T4 *patch the 8.333 kHz TTL SAMPLE CLOCK from the MASTER SIGNALS module to the* CLK *input of the PCM ENCODER module.*

T5 *connect the V_{in} input socket to ground of the variable DC module.*

T6 *connect the frame synchronization signal FS to the oscilloscope ext. synch. input.*

T7 *on CH1-A display the frame synchronization signal FS. Adjust the sweep speed to show three frame markers. These mark the* ***end*** *of each frame.*

T8 *on CH2-A display the CLK signal.*

T9 *record the number of clock periods per frame.*

Currently the analog input signal is zero volts (V_{in} is grounded). Before checking with the oscilloscope, consider what the PCM output signal might look like. Make a sketch of this signal, fully annotated. Then:

T10 *on CH2-B display the PCM DATA from the* PCM DATA *output socket.*

Except for the alternating pattern of '1' and '0' in the frame marker slot, you might have expected nothing else in the frame (all zeros), because the input analog signal is at zero volts. But you do not now the coding scheme.

There *is* an analog *input* signal to the encoder. It is of zero volts. This will have been coded into a 4-bit binary *output* number, which will appear in *each* frame. It need not be '0000'. The *same* number appears in *each* frame because the analog input is *constant*.

Your display should be similar to that of Figure 3 below, except that this shows five frames (too many frames on the oscilloscope display makes bit identification more difficult).

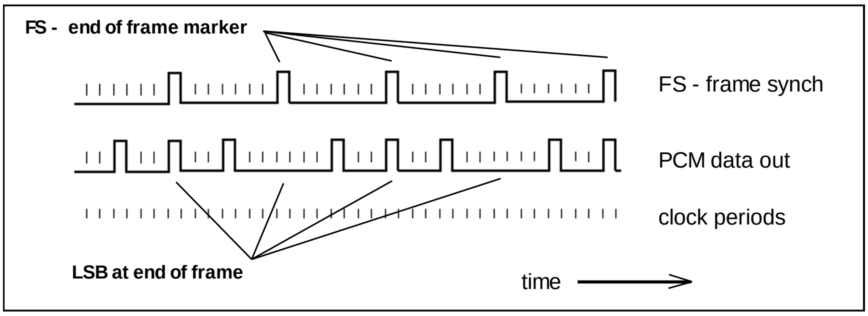


Figure 3: 5 frames of 4-bit PCM output for zero amplitude input

Knowing:

1. the number of slots per frame is 8
2. the location of the least significant bit is coincident with the end of the frame
3. the binary word length is four bits
4. the first three slots are 'empty' (in fact filled with zeros, but these remain unchanged under all conditions of the 4-bit coding scheme)

then:

T11 *identify the binary word in slots 4, 3, 2, and 1.*

quantizing levels for 4-bit linear encoding

You will now proceed to determine the quantizing/encoding scheme for the 4-bit linear case.

T12 *remove the ground connection, and connect the output of the VARIABLE DC module to V_{in}. Sweep the DC voltage slowly backwards and forwards over its complete range, and note how the data pattern changes in discrete jumps.*

T13 *if you have a WIDEBAND TRUE RMS METER module use this to monitor the DC amplitude at V_{in} - otherwise use the oscilloscope (CH1-B). Adjust V_{in} to its maximum negative value. Record the DC voltage and the pattern of the 4-bit binary number.*

T14 *slowly increase the amplitude of the DC input signal until there is a sudden change to the PCM output signal format. Record the format of the new digital word, and the input amplitude at which the change occurred.*

T15 *continue this process over the full range of the DC supply.*

T16 *draw a diagram showing the quantizing levels and their associated binary numbers.*

4-bit data format

From measurements made so far you should be able to answer the questions:

- what is the sampling rate ?
- what is the frame width ?
- what is the width of a data bit ?
- what is the width of a data word ?
- how many quantizing levels are there ?
- are the quantizing levels uniformly (linearly) spaced ?

7-bit linear encoding

T17 *change to 7-bit linear encoding by use of the front panel toggle switch.*

It would take a long time to repeat all of the above Tasks for the 7-bit encoding scheme. Instead:

T18 *make sufficient measurements so that you can answer all of the above questions in the section titled* ***4-bit data format*** *above. Making one or two assumptions (such as ?) you should be able to deduce the coding scheme used.*

companding

This module is to be used in conjunction with the PCM DECODER in a later experiment. As a pair they have a *companding* option. There is compression in the encoder, and expansion in the decoder. In the encoder this means the quantizing levels are closer together for small input amplitudes - that is, in effect, that the input amplitude peaks are compressed during encoding. At the decoder the 'reverse action' is introduced to restore an approximate linear input/output characteristic.

It can be shown that this sort of characteristic offers certain advantages, especially when the message has a high peak-to-average amplitude characteristic, as does speech, and where the signal-to-noise ratio is not high.

This improvement will not be checked in this experiment. But the existence of the non-linear quantization in the encoder will be confirmed.

In a later experiment, entitled *PCM decoding* (in this Volume), it will be possible to check the input/output linearity of the modules as a compatible pair.

T19 *change to 4-bit companding by use of the front panel toggle switch.*

T20 *the TIMS A_4 companding law has already been selected (first Task). Make the necessary measurements to determine the nature of the law.*

periodic messages

Although the experiment is substantially complete, you may have wondered why a periodic message was not chosen at any time. Try it.

T21 *take a periodic message from the SYNC. MESSAGE socket. This was set as the second Task.*

T22 *adjust the oscilloscope to display the message. Record its frequency and shape. Check if these are compatible with the Nyquist criterion; adjust the amplitude if necessary with one of the BUFFER AMPLIFIERS.*

T23 *now look at the PCM DATA output. Synchronize the oscilloscope (as previously) to the frame (FS) signal. Display two or three frames on CH1-A, and the PCM DATA output on CH2-A.*

You will see that the data signal reveals very little. It consists of many overlaid digital words, all different.

One would need more sophisticated equipment than is assumed here (a digital analyzer, a storage oscilloscope, ability to capture a single frame, and so on) to deduce the coding and quantizing scheme from such an input signal.

conclusions

What is the advantage of 7-bit over 4-bit encoding ? Of what use is companding ? From your measurements alone these questions cannot be answered.

These and other questions will be addressed in the experiment entitled *PCM decoding* (in this volume - but see the Tutorial Questions).

The findings of this experiment will be required in later PCM experiments. These will involve decoding of the data stream, an investigation of companding, and time multiplexing of the outputs from two PCM ENCODER modules.

Tutorial Questions

Q1 *from your knowledge of the PCM ENCODER module, obtained during preparation for the experiment, calculate the sampling rate of the analog input signal. Show that it is the same for both the 4-bit and the 7-bit coding schemes. What can you say about the bandwidth of an input analog signal to be encoded ?*

Q2 *define what is meant by the data 'frame' in this experiment. Draw a diagram showing the composition of a frame for:*

a) the 4-bit coding scheme

b) the 7-bit coding scheme

Q3 *it is possible to transmit each frame at a much slower rate than it was produced (and, of course, recover the original message as well). Explain how this might be done. When might this be an advantage ?*

Q4 *explain why a DC message gives a stable oscilloscope display of the PCM DATA output. Why is the display 'unstable' when a sine wave (for example) is the message ?*

Q5 *for the 4-bit encoder draw a diagram showing the amplitude quantization levels and the corresponding binary numbers used to encode them. Describe how this information was obtained experimentally.*

Q6 *two PCM signals can be combined to produce a time division multiplexed (PCM TDM) signal. With the measurements so far performed this does not seem (and indeed, is not) possible with two PCM ENCODER modules ! Why is this so ? Suggest what changes could be made to the module to implement PCM TDM [29].*

[29] in a later experiment it will be seen that suitable modifications to the data stream have been introduced so that a two-channel PCM TDM can be modelled.

Q7 *if you have studied the principles of companding in your course, describe its advantages. Then, if not already done so, plot the shape of the TIMS compression law introduced by the companding operation you measured. Compare this with published information about the 'A' and 'μ' companding laws used respectively in Europe and the USA.*

Appendix

For a MASTER CLOCK of 8.333 kHz, Table A-1 below gives the frequencies of the synchronized message at the SYNC. MESSAGE output for the setting of the on-board switch SW2.

For other clock frequencies the message frequency can be calculated by using the ‘divide by’ entry in the Table.

These messages are periodic, but not necessarily sinusoidal in shape. The term ‘sinuous’ means sine-like.

LH toggle	***RH toggle***	***divide clock by***	***freq with 8.333kHz clock***	***approx. ampl. and waveform***
UP	UP	32	260.4 Hz	0.2 V_{pp} sine
DOWN	UP	64	130.2 Hz	2.0 V_{pp} sine
UP	DOWN	128	65.1 Hz	4.0 V_{pp} sinuous
DOWN	DOWN	256	32.6 Hz	4.0 V_{pp} sinuous

Table A-1

9. PCM DECODING

ACHIEVEMENTS: *decoding of a PCM signal. Determination of the quantizing scheme used at the encoder. Message reconstruction. Introduction to companding; comparison of 7-bit linear with 4-bit companded PCM.*

PREREQUISITES: *completion of the experiment entitled* ***PCM encoding*** *in this Volume. An appreciation of the principles of companding.*

ADVANCED MODULES: *A PCM ENCODER and a PCM DECODER (version 2 preferable).*

Preparation

signal source

The signal to be decoded [30] in this experiment will be provided by you, using the PCM ENCODER module as set up in the experiment entitled *PCM encoding*. The format of the PCM signal is described there. You should have already completed that experiment.

clock synchronization

A clock synchronization signal will be stolen from the encoder.

frame synchronization

automatic

In the PCM DECODER module there is circuitry which automatically identifies the location of each frame in the serial data stream. To do this it collects groups of eight data bits and looks for the repeating pattern of alternate ones and zeros placed there (embedded) by the PCM ENCODER in the LSB position.

It can be shown that such a pattern cannot occur elsewhere in the data stream provided that the original bandlimited analog signal is sampled at or below the Nyquist rate.

When the embedded pattern is found an ‘end of frame’ synchronization signal FS is generated, and made available at the front panel.

The search for the frame is continuously updated. Why ?

Under noisy conditions (not relevant for this particular experiment) the reliability of the process will depend upon the size of the group of frames to be examined. This can be set by the on-board switch SW3 of the PCM DECODER module. See *Table A-1* in the Appendix to this experiment for details.

[30] it is common practice to refer to messages being *demodulated* from analog signals, and *decoded* from digital signals.

stolen

Frame synchronization can also be achieved, of course, by 'stealing' the synchronization signal, FS, from the PCM ENCODER module. Use of this signal would assume that the clock signal to the PCM DECODER is of the correct phase. This is assured in this experiment, but would need adjustment if the PCM signal is transmitted via a bandlimited channel (see Tutorial Question **7**). Hence the embedded frame synchronization information.

companding

You should prepare by reading something about the principles of *companding*. You will already be aware that the PCM ENCODER module can incorporate compression into its encoding scheme. The PCM DECODER module can introduce the complementary expansion. The *existence* of these characteristics will be confirmed, but their *effectiveness* in intelligibility enhancement (when speech is the message) is not examined.

PCM decoding

The PCM DECODER module is driven by an external clock. This clock signal is synchronized to that of the transmitter. For this experiment a 'stolen' clock will be used. The source of frame timing information has been discussed above.

Upon reception, the PCM DECODER:

1. extracts a frame synchronization signal FS from the data itself (from the embedded alternate ones and zeros in the LSB position), or uses an FS signal stolen from the transmitter (see above).
2. extracts the binary number, which is the coded (and quantized) amplitude of the sample from which it was derived, from the frame.
3. identifies the quantization level which this number represents.
4. generates a voltage proportional to this amplitude level.
5. presents this voltage to the output V_{out}. The voltage appears at V_{out} for the duration of the frame under examination.
6. message reconstruction can be achieved, albeit with some distortion, by lowpass filtering. A built-in reconstruction filter is provided in the module.

encoding

At the encoder the sample-and-hold operation (before encoding) is executed periodically. It produces a rectangular pulse form [31]. Each pulse in the waveform is of *exactly* the same amplitude as the message *at the sampling instant*.

But *it is not possible* to recover a *distortionless* message from these samples. They are *flat top*, rather than *natural* samples.

Call this the sampling distortion.

[31] if the sample is held for as long as the sampling period, it is a stepped waveform. If the sample is held for a shorter time it is a rectangular waveform (or pulseform). It need only be held long enough for the quantizer to make its decision about which of the available (quantized) amplitudes to allocate to the sample.

At the encoder the amplitude of this waveform was then *quantized.* It is still a rectangular pulsed waveform, but the amplitude of each pulse will, in general, be in error by a small amount. Call this waveform s(t).

This was examined in the experiment entitled *Sampling with SAMPLE & HOLD* (in this Volume), to which you should refer.

decoding

The voltage at V_{out} of the decoder is *identical with* s(t) above. The decoder itself has introduced no distortion of the received signal.

But s(t) is already an inexact version of the sample-and-hold operation at the encoder. This will give rise to *quantization distortion* as well as the *sampling distortion* already mentioned.

You should read about these phenomena in a Text book.

the TIMS PCM DECODER module

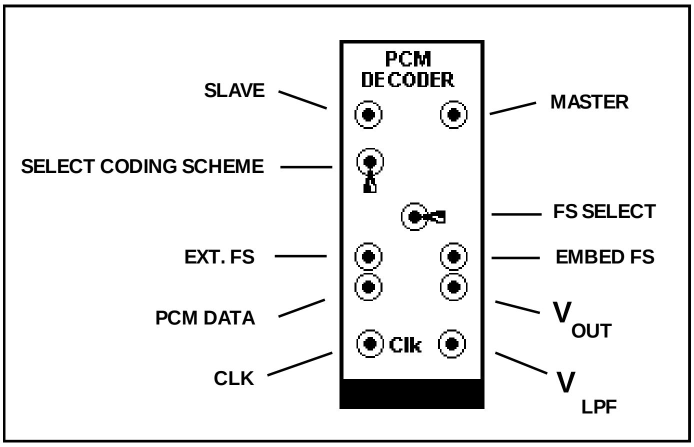


Figure 3: front panel layout of the PCM DECODER

A TIMS PCM DECODER module will be used for decoding.

The front panel of this module is shown in Figure 3. Technical details are described in the TIMS *Advanced Modules User Manual*.

Note and understand the purpose of the input and output connections, and the toggle switches. Counting from the top, these are:

- ***SLAVE***: not used during this experiment. Do *not* connect anything to this input.
- ***MASTER***: not used during this experiment. Do *not* connect anything to this output.
- ***SELECT CODING SCHEME***: a three position toggle which selects the coding scheme used by the signal to be decoded
- ***FS SELECT***: a two-position toggle switch which selects the method of obtaining the frame synchronization signal (FS) either external at (EXT.FS), or derived internally from the embedded information in the received PCM itself (EMBED FS).

- ***EXT. FS**:* connect an external frame sync. signal here if this method of frame synchronization is to be used.
- ***EMBED FS**:* if the frame synch. signal is derived internally from the embedded information, it is available for inspection at this output.
- ***PCM DATA**:* the PCM signal to be decoded is connected here.
- ***V_{OUT}**::* the decoded PCM signal.
- ***CLK**:* this is a TTL (red) input, and serves as the MASTER CLOCK for the module. Clock rate must be 10 kHz or less. For this experiment you will use the 8.333 kHz TTL signal from the MASTER SIGNALS module.

Experiment

the transmitter (encoder)

A suitable source of PCM signal will be generated using a PCM ENCODER module. This module was examined in the experiment entitled *PCM encoding*.

You should set it up before patching up the demodulator.

T1 *before plugging in PCM ENCODER module, set the toggles of the on-board SYNC MESSAGE switch SW2. Set the left hand toggle DOWN, and the right hand toggle UP. This selects a 130 Hz sinusoidal message, which will be used later. Now insert the module into the TIMS system.*

T2 *use the 8.333 kHz TTL signal from the MASTER SIGNALS module for the CLK.*

T3 *select, with the front panel toggle switch, the 4-bit LINEAR coding scheme.*

T4 *synchronize the oscilloscope 'externally' to the frame synchronization signal at FS. Set the sweep speed to 0.5 ms/cm (say). This should show a few frames on the screen.*

T5 *connect CH1-A of the SCOPE SELECTOR to the PCM OUTPUT of the PCM ENCODER.*

T6 *we would like to recognise the PCM DATA out signal. So choose a 'large' negative DC for the message (from the VARIABLE DC module). From previous work we know the corresponding code word is '0000', so only the embedded alternating '0' and '1' bits (for remote FS) in the LSB position should be seen. Confirm this. They should be 1920 ms apart. Confirm this both by measurement and calculation !*

T7 *vary the DC output and show the appearance of new patterns on CH1-A. When finished, return the DC to its maximum negative value (control fully anti-clockwise).*

The PCM signal is now ready for transmission. In a later experiment the PCM signal will be sent via a noisy, bandlimited channel. For the present it will be connected directly to a TIMS PCM DECODER module.

the receiver (decoder)

T8 *use the front panel toggle switch to select the 4-bit LINEAR decoding scheme (to match that of the transmitter)*

T9 *'steal' an 8.333 kHz TTL clock signal from the transmitter and connect it to the CLK input.*

T10 *in the first instance 'steal' the frame synchronization signal FS from the transmitter by connecting it to the frame synchronization input FS of the receiver. At the same time ensure that the FS SELECT toggle switch on the receiver is set to EXT. FS.*

T11 *ensure both channels of the oscilloscope are set to accept DC; set their gains to 1 volt/cm. With their inputs grounded set their traces in the centre of their respective halves of the screen. Remove the grounds.*

T12 *connect CH2-A to the sample-and-hold output of the PCM DECODER.*

!

a DC message

You are now ready to check the overall transmission from transmitter input to decoder output. The message is a DC signal.

T13 *connect the PCM DATA output signal from the transmitter to the PCM DATA input of the receiver.*

T14 *slowly vary the DC output from the VARIABLE DC module back and forth over its complete range. Observe the behaviour of the two traces. The input to the encoder moves continuously. The output from the decoder moves in discrete steps. These are the 16 amplitude quantizing steps of the PCM ENCODER.*

You are observing the source of quantizing noise. The output can take up only one of 16 predetermined values.

T15 *draw up a table relating input to output voltages.*

You can now see the number of quantizing levels at the transmitter, and their values.

T16 *compare the quantizing levels just measured with those determined in the experiment entitled* ***PCM encoding****.*

T17 *reset the coding scheme on both modules to 7-bit. Sweep the input DC signal over the complete range as before. Notice the 'granularity' in the output is almost un-noticeable compared with the 4-bit case. There are now 2^7 rather than 2^4 steps over the range.*

a periodic message

It was not possible, when examining the PCM ENCODER in the experiment entitled *PCM encoding*, to see the sample-and-hold waveform within the *encoder*. But you have just been looking at it (assuming perfect decoding) at the output of the *decoder*.

With a periodic message its appearance may be more familiar to you.

T18 *change to a periodic message* [32] *by connecting the* SYNC MESSAGE *of the PCM ENCODER,* ***via*** *a BUFFER AMPLIFIER, to its input V_{in}. An amplitude of 2 Vpp is suitable. Slow down the oscilloscope sweep speed to 1 ms/cm. Observe and record the signal at CH2-A.*

When you agree that what you see is what you expected to see, prepare to make a change and predict the outcome.

Currently the encoding scheme is generating a 4-bit digital word for each sample.

What would be the change to the waveform, now displaying on CH2-A, if, at the encoder, the coding scheme was changed from 4-bit to 7-bit ?

Sketch your answer to this question - show the waveform *before* and then *after* the change.

T19 *change the coding scheme from 4-bit to 7-bit. That is, change the front panel toggle switch of* ***both*** *the PCM ENCODER* ***and*** *the PCM DECODER from 4-bit to 7-bit. Observe, record, and explain the change to the waveform on CH2-A.*

When satisfied, proceed.

message reconstruction

You can see, qualitatively, that the output is related to the input. The message could probably be recovered from this waveform. But it would be difficult to predict with what accuracy.

Lowpass filtering of the waveform at the output of the decoder will reconstruct the message, although theory shows that it will not be perfect. It will improve with the number of quantizing levels.

What amplitude characteristic is required for the reconstruction filter ? See Tutorial Question Q3.

If any distortion components are present they would most likely include harmonics of the message. If these are to be measurable (visible on the oscilloscope, in the present case), then they must not be removed by the filter and so give a false indication of performance. Recall the experiment entitled *Amplifier overload* (within *Volume A2 - Further & Advanced Analog Experiments*).

[32] the message was set up in Task 1 to be a 130 Hz sinewave, synchronized to the sampling rate

So we could look for harmonics in the output of the filter. But we do not have conveniently available a spectrum analyzer.

An alternative is to use a two-tone test message. Changes to its shape (especially its envelope) are an indication of distortion, and are more easily observed (with an oscilloscope) than when a pure sinewave is used. It will be difficult to make one of these for this experiment, because our messages have been restricted to rather low frequencies, which are outside the range of most TIMS modules.

But there is provided in the PCM ENCODER a message with a shape slightly more complex than a sinewave. It can be selected with the switch SW2 on the encoder circuit board. Set the left hand toggle UP, and the right hand toggle DOWN. See the Appendix to the experiment entitled *PCM encoding* for more details.

A message reconstruction LPF is installed in the PCM DECODER module (version 2 and above). If you do not have such a module then bypass the next two Tasks.

T20 *change to the complex message from the PCM ENCODER as described above.*

T21 *include the built-in LPF in the output of the PCM DECODER, and observe the reconstructed message. Make comparisons between the 4-bit linear and the 7-bit linear coding schemes. Try different message amplitudes into the PCM ENCODER. Can you observe any distortion ? Record your observations.*

If you think the LPF itself might have introduced some distortion, you could check by connecting the complex message to its input direct, and observing the output.

companding

It is now time to verify the companding algorithm installed in the encoder.

T22 *use the front panel toggle switches (on both modules) to select 4-bit companding. Use both 'low' and 'high' level messages into the PCM ENCODER. Check the quantizing characteristic. Record your observations and comment upon them.*

Because of the speed limitations of the PCM modules it is not possible to use speech as the message, and so to observe the effects of companding. The effective bandwidth of the system is not wide enough.

See Tutorial Question Q7.

frame synchronization

In all of the above work the frame synchronization signal FS has been stolen from the encoder (as has been the clock signal). This was not necessary.

The PCM ENCODER has circuitry for doing this automatically. It looks for the alternating '0' and '1' pattern embedded as the LSB of each frame. It is enabled by use of the FS SELECT front panel toggle switch. Currently this is set to EXT FS.

T23 *change the FS SELECT switch on the front panel of the PCM DECODER module from EXT FS to EMBED. Notice that frame synchronization is re-established after a 'short time'. Could you put an upper limit on this time ? See Tutorial Question Q4.*

Tutorial Questions

Q1 *in the present experiment a 'stolen' clock signal was used. Why would transmission of the PCM signal via a bandlimited channel necessitate phase adjustment of this stolen clock signal to the PCM DECODER ?*

Q2 *sketch the waveforms at the output V_{out} from the decoder, for the 4-bit and the 7-bit linear encoding scheme (and a 'large amplitude' sinusoidal, synchronous message at the encoder). A 'sketch' might show these as being the same, but a more accurate drawing would show more clearly the difference. Explain.*

Q3 *how would you arrive at a specification for the reconstruction filter used in this experiment ?*

Q4 *from the information in Table A-1 make some quantitative comments on the length of time the built in circuitry of the PCM DECODER would take to recover the frame synchronization signal FS from the incoming data stream. Were you able to verify this by observation ?*

Q5 *two sources of distortion of the reconstructed message have been identified; they were called sampling distortion and quantizing distortion.*

a) assuming a sample-and-hold type sampler, what can be done about minimizing sampling distortion ?

b) what can be done about minimizing quantizing distortion ?

Q6 *quantizing distortion decreases with the number of quantizing levels available. There is usually a price to be paid for such an option. What would this be ? Was that apparent in the present experiment ? Explain.*

Q7 *companding is claimed to offer certain advantages. What are they ? Were you able to demonstrate any of these during the experiment. Explain.*

Appendix

automatic frame synchronization

The PCM DECODER module has built in circuitry for locating the position of each frame in the serial data stream. The circuitry looks for the embedded and alternating '0' and '1' in the LSB position of each frame.

The search is made by examining a section of data whose length is a multiple of eight bits.

The length of this section can be changed by the on-board switch SW3. Under noisy conditions it is advantageous to use longer lengths.

The switch settings are listed in Table A-1 below.

left toggle	right toggle	groups of eight bits
UP	UP	4
UP	DOWN	8
DOWN	UP	16
DOWN	DOWN	32

Table A-1: synchronization search length options

10. DELTA MODULATION

ACHIEVEMENTS: *an introduction to the basic delta modulator; to observe effects of step size and sampling clock rate change; slope overload and granular noise.*

PREREQUISITES: *some exposure to the principles of delta modulation in course work*

ADVANCED MODULES: *DELTA MODULATION UTILITIES, WIDEBAND TRUE RMS METER*

Preparation

principle of operation

Delta modulation was introduced in the 1940s as a simplified form of pulse code modulation (PCM), which required a difficult-to-implement analog-to-digital (A/D) converter.

The output of a delta modulator is a bit stream of samples, at a relatively high rate (eg, 100 kbit/s or more for a speech bandwidth of 4 kHz) the value of each bit being determined according as to whether the input message sample amplitude has increased or decreased relative to the previous sample. It is an example of differential pulse code modulation (DPCM).

block diagram

The operation of a delta modulator is to periodically sample the input message, to make a comparison of the current sample with that preceding it, and to output a single bit which indicates the sign of the difference between the two samples. This in principle would require a sample-and-hold type circuit.

De Jager (1952) hit on an idea for dispensing with the need for a sample and hold circuit. He reasoned that if the system *was* producing the desired output then this output could be sent back to the input and the two analog signals compared in a comparator. The output is a delayed version of the input, and so the comparison is in effect that of the current bit with the previous bit, as required by the delta modulation principle.

Figure 1 illustrates the basic system in block diagram form, and this will be the modulator you will be modelling.

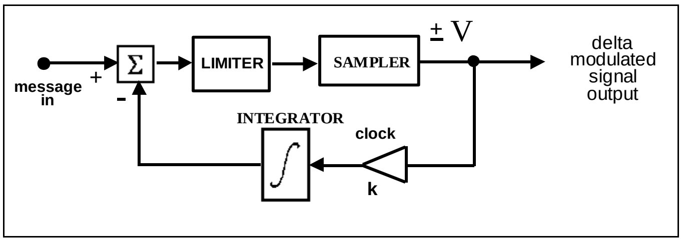


Figure 1: basic delta modulator

The system is in the form of a feedback loop. This means that its operation is not necessarily obvious, and its analysis non-trivial. But you can build it, and confirm that it does behave in the manner a delta modulator should.

The system is a continuous time to discrete time converter. In fact, it is a form of analog to digital converter, and is the starting point from which more sophisticated delta modulators can be developed.

The sampler block is clocked. The output from the sampler is a bipolar signal, in the block diagram being either ±V volts. This is the delta modulated signal, the waveform of which is shown in Figure 2. It is fed back, in a feedback loop, via an integrator, to a summer.

The integrator output is a sawtooth-like waveform, also illustrated in Figure 2. It is shown overlaid upon the message, of which it is an approximation.

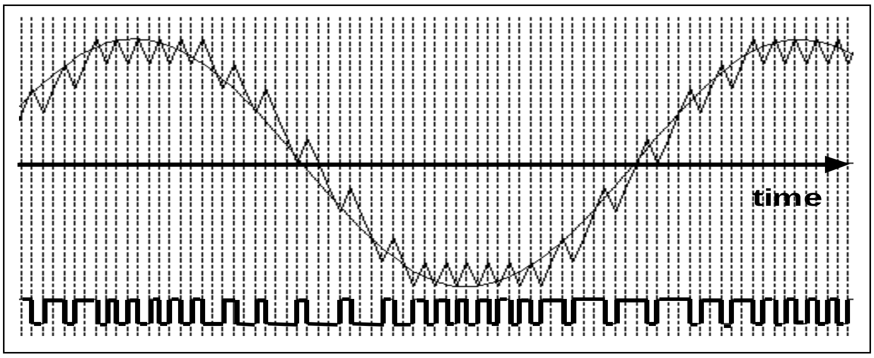


Figure 2: integrator output superimposed on the message with the delta modulated signal below

The sawtooth waveform is subtracted from the message, also connected to the summer, and the difference - an error signal - is the signal appearing at the summer output.

An amplifier is shown in the feedback loop.. This controls the loop gain. In practice it may be a separate amplifier, part of the integrator, or within the summer. It is used to control the size of the ‘teeth’ of the sawtooth waveform, in conjunction with the integrator time constant.

When analysing the block diagram of Figure 1 it is convenient to think of the summer having unity gain between both inputs and the output. The message comes in at a fixed amplitude. The signal from the integrator, which is a sawtooth approximation to the message, is adjusted with the amplifier to match it as closely as possible. You will be able to see this when you make a model of the system of Figure 1.

step size calculation

In the delta modulator of Figure 1 the output of the integrator is a sawtooth-like approximation to the input message. The teeth of the saw must be able to rise (or fall) fast enough to follow the message. Thus the integrator time constant is an important parameter.

For a given sampling (clock) rate the step *slope* (volt/s) determines the *size* (volts) of the step within the sampling interval.

Suppose the amplitude of the rectangular wave from the sampler is ±V volt. For a change of input sample to the integrator from (say) negative to positive, the change of integrator output will be, after a clock period T:

$$output = \frac{2kVT}{RC} \ \ volt \qquad \text{........ 34}$$

where ***k*** is the gain of the amplifier preceding the integrator (as in Figure 1).

Answer Tutorial Questions 1 and 2 before attempting the experiment. You can later check your answer by measurement.

slope overload and granularity

The binary waveform illustrated in Figure 2 is the signal transmitted. This is the delta modulated signal.

The integral of the binary waveform is the sawtooth approximation to the message.

In the experiment entitled *Delta demodulation* (in this Volume) you will see that this sawtooth wave is the primary output from the demodulator at the receiver.

Lowpass filtering of the sawtooth (from the demodulator) gives a better approximation to the message. But there will be accompanying noise and distortion, products of the approximation process at the modulator.

The unwanted products of the modulation process, observed at the receiver, are of two kinds. These are due to 'slope overload', and 'granularity'.

slope overload

This occurs when the sawtooth approximation cannot keep up with the rate-of-change of the input signal in the regions of greatest slope.

The step size is reasonable for those sections of the sampled waveform of small slope, but the approximation is poor elsewhere. This is 'slope overload', due to too small a step.

Slope overload is illustrated in Figure 3.

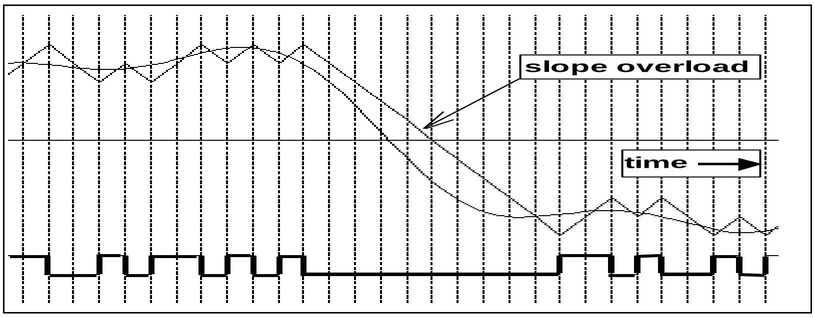


Figure 3: slope overload

To reduce the possibility of slope overload the step size can be increased (for the same sampling rate). This is illustrated in Figure 4. The sawtooth is better able to match the message in the regions of steep slope.

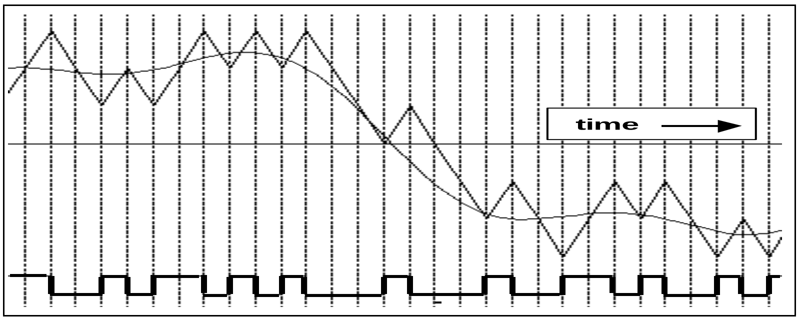


Figure 4: increased step size to reduce slope overload

An alternative method of slope overload reduction is to increase the sampling rate. This is illustrated in Figure 5, where the rate has been increased by a factor of 2.4 times, but the step is the same size as in Figure 3.

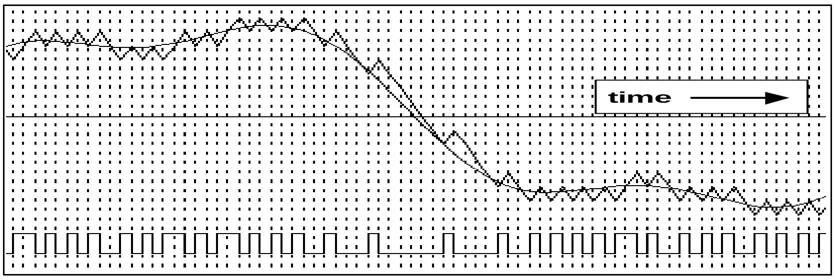


Figure 5: increased sampling rate to reduce slope overload

granular noise

Refer back to Figure 3. The sawtooth follows the message being sampled quite well in the regions of small slope. To reduce the slope overload the step size is increased, and now (Figure 4) the match over the regions of small slope has been degraded.

The degradation shows up, at the demodulator, as increased quantizing noise, or ‘granularity’.

noise and distortion minimization

There is a conflict between the requirements for minimization of slope overload and the granular noise. The one requires an increased step size, the other a reduced step size. You should refer to your text book for more discussion of ways and means of reaching a compromise. You will meet an example in the experiment entitled *Adaptive delta modulation* (in this Volume).

An optimum step can be determined by minimizing the quantizing error at the summer output, or the distortion at the demodulator output.

Experiment

The block diagram of Figure 1 is modelled with a DELTA MODULATION UTILITIES module, an ADDER, and both of the BUFFER AMPLIFIERS.

You should obtain a DELTA MODULATION UTILITIES module, and read about it in the *TIMS Advanced Modules User Manual*. This module contains three of the elements of the block diagram, namely the LIMITER, SAMPLER, and INTEGRATOR.

The SUMMER block is modelled with an ADDER, both gains being set to unity.

The amplifier preceding the INTEGRATOR in the feedback loop is modelled by a *pair* of BUFFER AMPLIFIERS connected in cascade. These amplifiers both invert, so the combination will be non-inverting - as required.

If the ADDER gains are left fixed at unity, and the message and sampling rates fixed, the only variables to be investigated are the INTEGRATOR time constant, and the gain ***k*** of the amplifier (the two BUFFERS in cascade) in the feed back loop.

setting up

T1 *obtain and examine a DELTA MODULATOR UTILITIES module. Read about it in the **TIMS Advanced Modules User Manual**. Before plugging it in set the on-board switches to give an intermediate INTEGRATOR time constant (say SW2A to ON, and SW2B to OFF). Start with no division of the 100 kHz sample clock (front panel toggle switch up to 'CLK').*

T2 *plug in the ADDER and DELTA MODULATION UTILITIES module.*

T3 *use a sinewave to set both of the ADDER gains close to unity. **Do not change these for the duration of the experiment.***

T4 *use a sinewave to set both of the BUFFER AMPLIFIER gains to about unity (they are connected in series to make a non-inverting amplifier). Either one or both of these will be varied to make adjustments to the step size during the course of the experiment.*

T5 *patch up a model of Figure 1. This is shown in Figure 6. Use the 100 kHz TTL signal from the MASTER SIGNALS module as the clock for the SAMPLER, and the 2 kHz MESSAGE for the sinusoidal message to be sampled. The message (2.083 kHz) is a sub-multiple of the 100 kHz sample clock. This helps to obtain text-book like oscilloscope displays.*

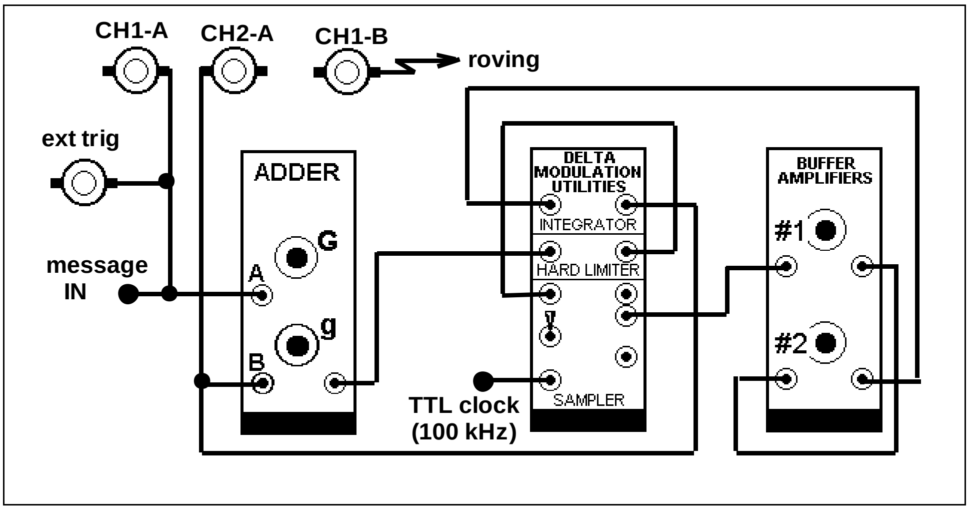


Figure 6: the delta modulator; a model of Figure 1

T6 *use the 2 kHz message as the 'ext. trig' signal to the oscilloscope. The signals of immediate interest are the two inputs to the* SUMMER*, shown connected to CH1-A and CH2-A. Use CH1-B to explore other signals.*

You will now set up the modulator for 'acceptable performance'. This means that the INTEGRATOR output should be a reasonable approximation to the message at the input to the SUMMER (of Figure 1).

The *only adjustments you should make* during the course of the experiment are to:

1. ***the step size***: this can be varied in fixed steps with the INTEGRATOR time constant, or fine steps with the gain ***k*** of the amplifier (two cascaded BUFFER amplifiers) in the feedback loop (Figure 1).
2. ***the sampling clock rate***: with the front panel toggle switch of the DELTA MODULATOR UTILITIES module (100, 50, or 25 kHz).

You should keep a record of the waveforms observed. Sketch them all relative to the sampling clock. Make sure each is consistent with expectations before proceeding to the next Task.

Remember the ADDER is modelling the SUMMER (of Figure 1). The two inputs are the message and its approximation. These two should be of the same general shape and the same amplitude. Since it is their difference which is being sought they will need to be of opposite polarity, as has been arranged (remember, the gains **g** and **G** of the ADDER, acting as a SUMMER, have both been set to unity).

Observe the two inputs to the SUMMER. You should use the 'inverse' facility of your oscilloscope (or one channel of another ADDER set to unity gain) and overlay the two displays to simplify their comparison.

> ***warning***: *remember, when recording other observations, to restore the inverse operation of the oscilloscope to normal.*

T7 *examine the two inputs to the ADDER on CH1-A and CH2-A. These are the input message, and the* INTEGRATOR *output respectively. Remember that the* INTEGRATOR *waveform is required to be an approximation to the message. Adjust the gain **k** to achieve what you consider the 'best' match. You should have a display similar to that of Figure 2.*

You will notice that, despite the fact that the message is a sub-multiple of the clock rate, it is also necessary to fine-tune the oscilloscope sweep speed to obtain a totally stable oscilloscope display. This is through no fault of the oscilloscope - think about it !

T8 *find and measure the smallest amplitude step between samples in the* INTEGRATOR *output waveform over a single clock period. This is the quantizing interval, or step size. Observe that larger steps occur over more than one clock period and that small steps occur when the rate of change of the input is small (near the extrema of the sinewave message). Verify, by calculation, the step size.*

Describe in your notes what happens to the approximation when ***k*** is decreased, and when ***k*** is increased.

T9 *observe the ADDER output and confirm that it is the difference between, rather than the sum of, the two inputs. This is the quantizing noise (quantizing error). Notice that not all peaks are of the same height - there are occasional large peaks. Use the WIDEBAND TRUE RMS METER to measure the quantizing noise (remove any DC with the front panel switch). Adjust the step size with the gain **k** to minimize the quantizing error. Measure the peak-to-peak amplitude, and rms amplitude. Compare with theoretical expectations. Refer Tutorial Question Q4.*

You will have a chance to measure the distortion of the demodulated signal in the experiment entitled *Delta demodulation* in this Volume. The amount of distortion can be used as another *quantitative* criterion for setting ***k***.

slope overload

The adjustment of the gain ***k***, as a means of controlling slope overload, has so far been made while watching the INTEGRATOR output. This is a *qualitative* judgement of slope overload.

the ADAPTIVE CONTROL *signal*

In terms of the principle of operation of the delta modulator slope overload gives rise to a succession of samples from the SAMPLER module *of the same sign*. This condition can be detected electronically.

The DELTA MODULATION UTILITIES module has such detection circuitry. When three or more consecutive samples are of the same sign this circuitry signals the fact with a +4 volt output from the ADAPTIVE CONTROL socket of the SAMPLER module. Otherwise the output is at a level of about +2 volt. This signal is used in a later experiment (entitled *Adaptive delta modulation* in this Volume); for now it is instructive to monitor it, for an independent (and more reliable ?) indication of slope overload.

T10 *vary the gain **k**, and watch the INTEGRATOR output (CH2-A) for signs of slope overload; at the same time monitor the ADAPTIVE CONTROL signal (CH1-B) and compare its pronouncement with your judgement. Since this is a time-sensitive (phase) measurement, make sure your oscilloscope is set up correctly (eg, not on 'alternate-trace' mode). Record how many clock periods elapse, following the onset of slope overload, before this is signalled by the ADAPTIVE CONTROL output signal.*

T11 *re-adjust for 'moderate' slope overload. Increase and decrease the step size by means of the INTEGRATOR time constant (SW2A and SW2B on the DELTA MODULATION UTILITIES module circuit board). Confirm that the degree of slope overload changes as expected.*

T12 *the front panel switch of the DELTA MODULATION UTILITIES module inserts dividers between the clock input and the SAMPLER, to vary the clock rate. Select an intermediate clock rate, and re-adjust for 'moderate' slope overload. Show that slope overload increases when the clock speed is halved, or decreases when the clock rate is doubled. Does the step size change when the clock changes ?*

the output

So far you have not looked at the output signal from the modulator ! Generally this is the first signal to look at.

The output signal is in TTL format. It is a HI if the INTEGRATOR output is rising, and a LO otherwise. The output signal appears in each of Figures 2,3, 4 and 5.

T13 *use CH1-B to look at the modulator output - that is, from the SAMPLER. Compare it with the INTEGRATOR output on CH2-A. Confirm the relationship between the two waveforms.*

T14 *observe the relationship between the delta modulator output (CH1-B) and the clock signal (use CH2-B).*

So far, as promised, there were only two parameters to be varied during the course of the experiment - the loop gain factor **k**, and the integrator time constant. These were sufficient to allow many observations to be made.

If all of the above has been appreciated it might be a good idea to predict what might happen if the message frequency was changed. Consider the possibilities, then make the change.

T15 *set up as for the conditions of Task T7 (whilst observing the two inputs to the SUMMER).*

T16 *set an AUDIO OSCILLATOR to about 2 kHz, and use it for the message (and ext trig signal), instead of the synchronous 2.083 kHz message. Leaving all other variables fixed, vary the message frequency. Whilst it is not easy to stabilise the display, it is still possible to see some consequences, including the onset of slope-overload. Record and explain your observations.*

complex message

A sinewave message is useful for many tests, but a more complex shape can lead to more insights. For meaningful oscilloscope displays it will need to be periodic, and, as before, a sub-multiple of the sampling rate.

Such a message is easy to make with TIMS; a possible method is described in the Appendix to this experiment.

Such a message was used to produce the waveforms of Figures 3, 4, and 5.

T17 *make a synchronous, complex message. Vary its shape, and observe results under different conditions.*

Tutorial Questions

Q1 *why is it useful to set up the experiment using the 2 kHz signal from the MASTER SIGNALS module, as opposed to a signal from an AUDIO OSCILLATOR, for example ?*

Q2 *what are the system parameters which control the step size (quantization amplitude) for a given sampling rate ?*

Q3 *knowledge of the step size alone is insufficient to make a statement about the possibility of slope overload. What else needs to be known ?*

Q4 *calculate the peak-to-rms ratio of a constant peak-to-peak amplitude sawtooth waveform.*

Q5 *show that delta modulation is a special case of differential pulse code modulation (DPCM). What is the number of bits per word ?*

Appendix

a 'complex' message.

We can define a 'complex' message as one which is periodic, and having a shape exhibiting more slope changes than a pure sinewave. As an example, see Figure A.1.

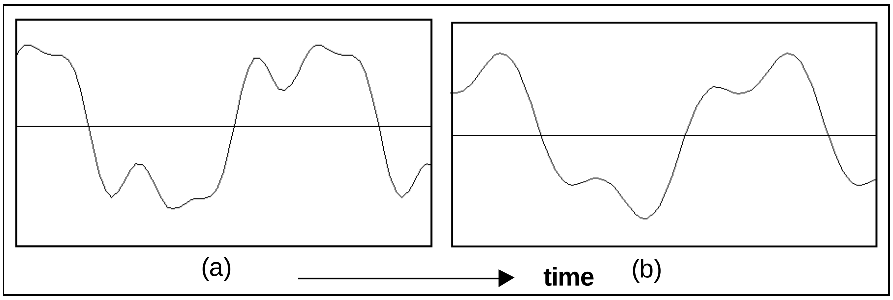


Figure A1: two 'complex' messages

Such a message can be made by filtering a square wave.

The square wave can be obtained by passing a sinewave through a comparator. The TIMS COMPARATOR has an analog output. Its limiting characteristic can be set to 'hard' [33]. See the ***TIMS User Manual***. The amplitude limited output contains odd harmonics, and the first two or three can be filtered off (together with the fundamental) to make the new shape.

By including a PHASE SHIFTER (this introduces a phase shift which varies with frequency), the shape can be further modified; but this is not essential.

For synchronous displays, in the present experiment, it is useful to use the 2.083 kHz MESSAGE from MASTER SIGNALS.

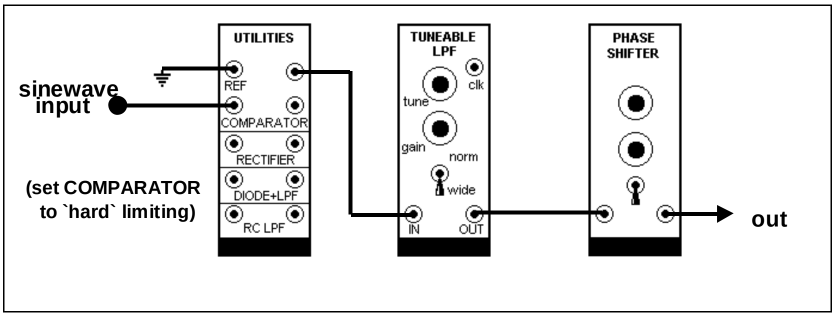


Figure A2: a 'complex' message generator

The waveforms of Figure A1 were made by selecting the first three odd harmonics (a), and the first two odd harmonics (b) respectively. Many shape variations are possible, including these, as the phase is varied.

An interesting feature is that, by obtaining the complex waveform shape with the TUNEABLE LPF in the WIDE mode, instant reversion to a sine wave is effected by toggling to the NORM mode.

[33] SW1 toggles DOWN; SW2 toggles UP

11.DELTA DEMODULATION

ACHIEVEMENTS: *introduction to the demodulation of a delta modulated signal; measurement of quantization distortion at the receiver; listening test on speech.*

PREREQUISITES: *completion of the experiment entitled* **Delta modulation** *in this Volume.*

ADVANCED MODULES: *DELTA MODULATION UTILITIES; DELTA DEMOD UTILITIES; WIDEBAND TRUE RMS METER*

Preparation

delta demodulation methods

You should refer to your text book and course work for background information regarding delta demodulation methods, and the likely sources of distortion.

For this experiment you will supply your own delta modulated signal, using the modulator examined in the experiment entitled *Delta modulation*.

The TIMS DELTA DEMOD UTILITIES module will be used for demodulation (the receiver). It contains a SAMPLER and an INTEGRATOR. The SAMPLER uses a clock stolen from the modulator (the transmitter). The SAMPLER accepts TTL signals as input, but gives an analog output for further analog processing - for example, lowpass filtering.

The principle of the demodulator is shown in block diagram form in Figure 1 below. It performs the reverse of the process implemented at the modulator in the vicinity of the sampler and integrator.

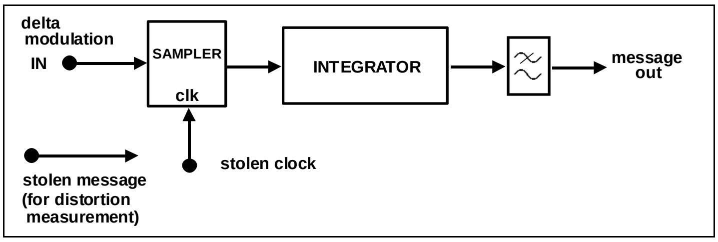


Figure 1: a demodulator for delta modulation

The sampler, which is clocked at the same rate as the one at the modulator, outputs a bi-polar signal (±V volts). The integrator generates a sawtooth-like waveform from this. This is an approximation to the original message. Having the same time constant as that at the modulator, and with no noise or other signal impairments, it will be identical with the corresponding signal at the modulator.

However, it is not the message, but an approximation to it.

The sawtooth waveform contains information at the message frequency, plus obvious unwanted frequency components (quantizing noise).

The unwanted components which are beyond the bandwidth of the original baseband message are removed by a lowpass filter. Those unwanted components which remain are perceived as noise and distortion.

Unlike ideal sampling of an analog signal, and ideal reconstruction with a lowpass filter (refer to the experiment entitled *The sampling theorem* within *Volume A1 - Fundamental Analog Experiments*), the reconstruction of the message from a delta modulator is *not* perfect.

You will find that the SNDR [34] is relatively poor, and certainly a lot worse than the signal-to-noise ratio capabilities of the TIMS system (typically better than 40 dB). Thus the SNDR that you will be measuring will be entirely due to the imperfections of the delta modulator itself.

However, do not then declare that delta modulation has no practical applications.

You will find, in the experiment entitled *Adaptive delta modulation*, in this Volume, that there are means of implementing improvements.

With further refinement in the circuitry, a higher clock speed, and sophisticated adaptive algorithms [35], delta modulation can perform remarkably well. It is used extensively in the field of digital audio.

Experiment

test signal

T1 *set up a delta modulator of the type examined in the experiment entitled* **Delta modulation**. *Set it up initially for what you consider to be the 'best' approximation to the message (compare the two inputs to the* SUMMER*).*

the demodulator

For this demodulator you will use the DELTA DEMOD UTILITIES module. This contains a LIMITER, a clocked SAMPLER, and an INTEGRATOR of the type in the DELTA MODULATION UTILITIES module.

T2 *obtain and examine a DELTA DEMOD UTILITIES module. Read about it in the* ***TIMS Advanced Modules User Manual.***

34 signal-plus-noise-and-distortion ratio

35 see the experiment entitled *Adaptive delta modulation* in this Volume

T3 *model the demodulator of Figure 1. Set the time constant of the* INTEGRATOR *to the same value as selected in the modulator. Use the* RC LPF *in the DELTA DEMOD UTILITIES for the output filter.*

Note the SAMPLER *accepts a TTL signal from the modulator, as well as a stolen clock. For oscilloscope triggering use the message signal, also stolen from the modulator. Set the front panel clock switch to match that at the modulator.*

T4 *confirm that the signals at each of the* INTEGRATOR *outputs are similar.*

T5 *confirm that the output of the demodulator lowpass filter is a reasonable copy of the original message.*

distortion - a qualitative look

At the modulator you can change the sampling rate (100 kHz, 50 kHz, and 25 kHz with the front panel switch), and the step size (RC time constants). You can also control the amount of slope overload. All of these have their influence on the measured SNDR.

T6 *introduce various mal-adjustments at the modulator (observed at the output of the modulator* INTEGRATOR*), and observe their effect at the demodulator output. Use both a sinusoidal message, and a 'complex' message* [36]*.*

speech

If you have bandlimited speech available at TRUNKS, or from a SPEECH module, you can make many interesting listening tests. How would you describe speech when distorted by slope overload ?

T7 *make qualitative assessments of the effect of the various mal-adjustments, at the modulator, on the demodulated speech.*

distortion - SNDR measurement

For quantitative signal-to-noise-and-distortion ratio measurements (SNDR) you can model the scheme illustrated in Figure 2.

36 as defined in the experiment entitled *Delta modulation* in this Volume.

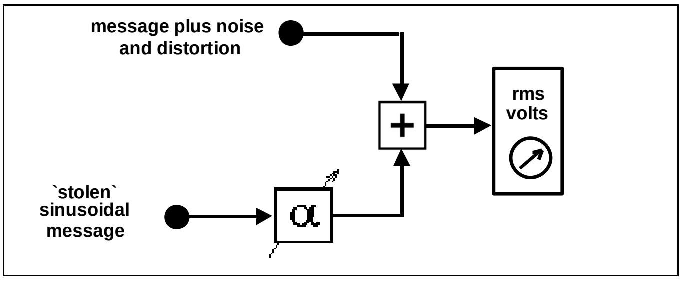


Figure 2: noise and distortion measurement

Recall the experiment entitled *Modelling an equation* (within *Volume A1 - Fundamental Analog Experiments*), where the technique of signal cancellation in an adder was first introduced.

You can use the WIDEBAND TRUE RMS METER to measure the distortion components.

The principle is to cancel the wanted sinusoidal message from the adder output, leaving only the unwanted components (noise-plus-distortion). Having obtained a minimization of the message from the adder output, then removal of the message-plus-noise from the adder leaves the (stolen) message, which will give the reference with which to compare the noise-plus-distortion.

A model of the measurement system is illustrated in Figure 3. Remember to set the on-board switch of the PHASE SHIFTER to LO.

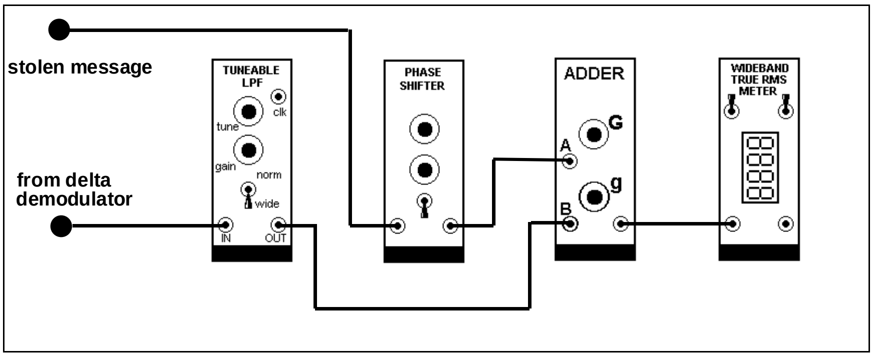


Figure 3: SNDR measurement

T8 *as before (when making qualitative observations), introduce various mal-adjustments at the modulator, and observe their effect at the demodulator output. Use a sinusoidal message (refer to Tutorial Question Q4).*

Tutorial Questions

Q1 *the term 'granular noise' is often used in the context of delta modulation. Explain where this term comes from. Describe the compromise which has to be made when determining a step size in a delta modulator.*

Q2 *describe the procedure used when measuring SNDR with the scheme of Figure 2.*

Q3 *what was the effect upon output noise and distortion of an increase of:*

a) step size

b) sampling rate

c) slope overload ?

Q4 *the noise-and-distortion measurement scheme of Figure 2 was used when the message was a single sine wave. Would it be effective for measurement with a more complex message ? Explain.*

Q5 *you were advised to set the time constant of the* INTEGRATOR *in the demodulator to be the same as that in the modulator. Was this essential ? Describe the consequences of using a different time constant at the demodulator.*

Q6 *could the message be recovered from the delta modulated signal using only a lowpass filter ? Explain.*

12. LINE CODING

ACHIEVEMENTS: *familiarity with the properties of the LINE-CODE ENCODER and LINE-CODE DECODER modules, and the codes they generate.*

PREREQUISITES: *an appreciation of the purpose behind line coding.*

EXTRA MODULES: *LINE-CODE ENCODER and LINE-CODE DECODER*

Preparation

This 'experiment' is tutorial in nature, and serves to introduce two new modules.

In your course work you should have covered the topic of line coding at what ever level is appropriate for you. TIMS has a pair of modules, one of which can perform a number of line code transformations on a binary TTL sequence. The other performs decoding.

why line coding ?

There are many reasons for using line coding. Each of the line codes you will be examining offers one or more of the following advantages:

spectrum shaping and relocation without modulation or filtering. This is important in telephone line applications, for example, where the transfer characteristic has heavy attenuation below 300 Hz.

bit clock recovery can be simplified.

DC component can be eliminated; this allows AC (capacitor or transformer) coupling between stages (as in telephone lines). Can control baseline wander (baseline wander shifts the position of the signal waveform relative to the detector threshold and leads to severe erosion of noise margin).

error detection capabilities.

bandwidth usage; the possibility of transmitting at a higher rate than other schemes over the same bandwidth.

At the very least the LINE-CODE ENCODER serves as an interface between the TTL level signals of the transmitter and those of the analog channel. Likewise, the LINE-CODE DECODER serves as an interface between the analog signals of the channel and the TTL level signals required by the digital receiver.

the modules

The two new modules to be introduced are the LINE-CODE ENCODER and the LINE-CODE DECODER.

You will not be concerned with how the coding and decoding is performed.

You should examine the waveforms, using the original TTL sequence as a reference.

In a digital transmission system line encoding is the final digital processing performed on the signal before it is connected to the analog channel, although there may be simultaneous bandlimiting and wave shaping.

Thus in TIMS the LINE-CODE ENCODER accepts a TTL input, and the output is suitable for transmission via an analog channel.

At the channel output is a signal at the TIMS ANALOG REFERENCE LEVEL, or less. It could be corrupted by noise. Here it is re-generated by a *detector*. The TIMS detector is the DECISION MAKER module (already examined in the experiment entitled *Detection with the DECISION MAKER* in this Volume). Finally the TIMS LINE-CODE DECODER module accepts the output from the DECISION MAKER and decodes it back to the binary TTL format.

Preceding the line code encoder may be a source encoder with a matching decoder at the receiver. These are included in the block diagram of Figure 1, which is of a typical baseband digital transmission system. It shows the disposition of the LINE-CODE ENCODER and LINE-CODE DECODER. All bandlimiting is shown concentrated in the channel itself, but could be distributed between the transmitter, channel, and receiver.

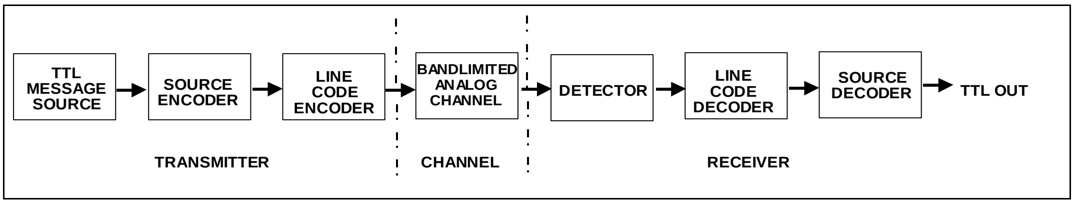


Figure 1: baseband transmission system

The LINE-CODE ENCODER serves as a source of the system bit clock. It is driven by a *master clock* at 8.333 kHz (from the TIMS MASTER SIGNALS module). It divides this by a factor of four, in order to derive some necessary internal timing signals at a rate of 2.083 kHz. This then becomes a convenient source of a 2.083 kHz TTL signal for use as the *system bit clock*.

Because the LINE-CODE DECODER has some processing to do, it introduces a time delay. To allow for this, it provides a re-timed clock if required by any further digital processing circuits (eg, for decoding, or error counting modules).

terminology

- the word *mark*, and its converse *space*, often appear in a description of a binary waveform. This is an historical reference to the mark and space of the telegraphist. In modern day digital terminology these have become HI and LO, or '1' and '0', as appropriate.
- *unipolar signalling*: where a '1' is represented with a finite voltage V volts, and a '0' with zero voltage. This seems to be a generally agreed-to definition.
- those who treat *polar* and *bipolar* as identical define these as signalling where a '1' is sent as +V, and '0' as -V. They append AMI when referring to three-level signals which use +V and -V alternately for a '1', and zero for '0' (an alternative name is pseudoternary).

You will see the above usage in the *TIMS Advanced Modules User Manual*, as well as in this text.

However, others make a distinction. Thus:

- *polar signalling*: where a '1' is represented with a finite voltage +V volts, and a '0' with -V volts.
- *bipolar signalling*: where a '1' is represented alternately by +V and -V, and a '0' by zero voltage.

- the term 'RZ' is an abbreviation of 'return to zero'. This implies that the particular waveform will return to zero for a finite part of each data '1' (typically half the interval). The term 'NRZ' is an abbreviation for 'non-return to zero', and this waveform will not return to zero during the bit interval representing a data '1'.
- the use of 'L' and 'M' would seem to be somewhat illogical (or inconsistent) with each other. For example, see how your text book justifies the use of the 'L' and the 'M' in NRZ-L and NRZ-M.
- two sinusoids are said to be antipodal if they are 180^0 out of phase.

available line codes

For a TTL input signal the following output formats are available from the LINE-CODE ENCODER.

NRZ-L

Non return to zero - level (bipolar): this is a simple scale and level shift of the input TTL waveform.

NRZ-M

Non return to zero - mark (bipolar): there is a transition at the beginning of each '1', and no change for a '0'. The 'M' refers to 'inversion on mark'. This is a differential code. The decoder will give the correct output independently of the polarity of the input.

UNI-RZ

Uni-polar - return to zero (uni-polar): there is a half-width output pulse if the input is a ‘1’; no output if the input is a ‘0’. This waveform has a significant DC component.

BIP-RZ

Bipolar return to zero (3-level): there is a half-width +ve output pulse if the input is a ‘1’; or a half-width -ve output pulse if the input is a ‘0’. There is a return-to-zero for the second half of each bit period.

RZ-AMI

Return to zero - alternate mark inversion (3-level): there is a half-width output pulse if the input is a ‘1’; no output if the input is a ‘0’. This would be the same as UNI-RZ. But, *in addition*, there is a polarity inversion of every alternate output pulse.

Biϕ-L

Biphase - level (Manchester): bipolar ±V volts. For each input ‘1’ there is a transition from +V to -V in the middle of the bit-period. For each input ‘0’ there is a transition from -V to +V in the middle of the bit period.

DICODE-NRZ

Di-code non-return to zero (3-level): for each transition of the input there is an output pulse, of opposite polarity from the preceding pulse. For no transition between input pulses there is no output.

The codes offered by the line-code encoder are illustrated in Figure 2 below. These have been copied from the *Advanced Module Users Manual*, where more detail is provided.

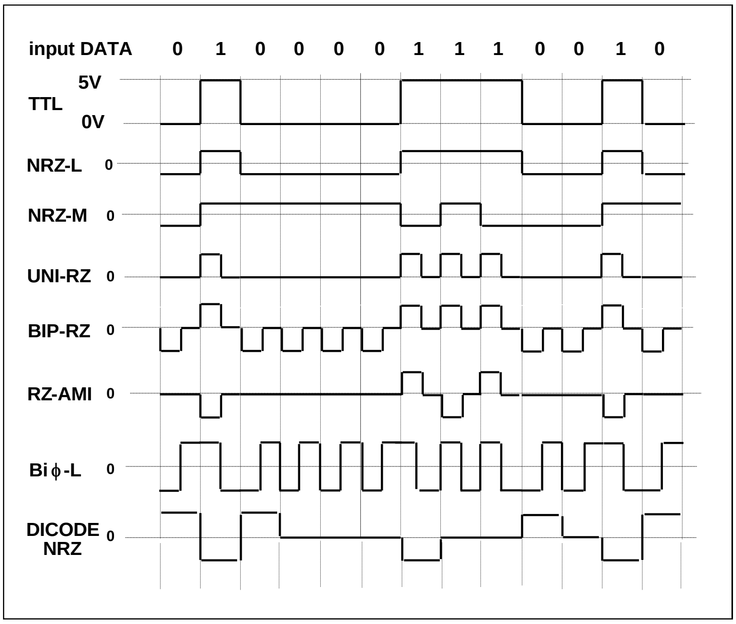


Figure 2: TIMS line codes

The output waveforms, apart from being encoded, have all had their amplitudes adjusted to suit a TIMS analog channel (not explicitly shown in Figure 2).

When connected to the input of the LINE-CODE DECODER these waveforms are de-coded back to the original TTL sequence.

band limiting

No matter what the line code in use, it is not uncommon to bandlimit these waveforms before they are sent to line, or used to modulate a carrier.

As soon as bandlimiting is invoked individual pulses will spread out (in the time domain) and interfere with adjacent pulses. This raises the issue if inter-symbol interference (ISI).

A study of ISI is outside the intended scope of this text, but it cannot be ignored in practice. Bandlimiting (by pulse shaping) can be effected and ISI controlled by appropriate filter design.

An alternative approach, duobinary encoding, was invented by Lender [37].

duobinary encoding

A duobinary encoder (and decoder) is included in the line code modules.

Duobinary encoding is also called correlative coding, or partial response signalling.

The precoded duobinary encoding model implemented in the LINE-CODE ENCODER module is described in the *TIMS Advanced Modules User Manual*.

---

[37] Lender, A. "The Duobinary Technique for High Speed Data Transmission", IEEE Trans. Comm. Electron, vol 82, pp. 214-218, May 1963

Experiment

Figure 3 shows a simplified model of Figure 1. There is no source encoding or decoding, no baseband channel, and no detection. For the purpose of the experiment this is sufficient to confirm the operation of the line code modules.

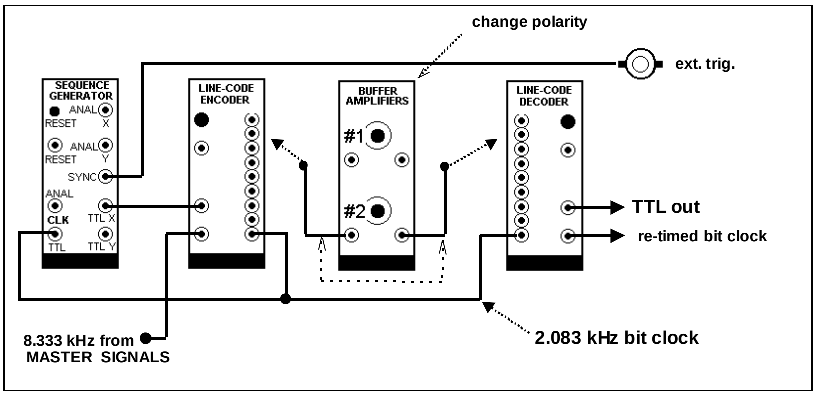


Figure 3: simplified model of Figure 1

When a particular code has been set up, and the message successfully decoded without error, the BUFFER should be included in the transmission path. By patching it in or out it will introduce a polarity change in the channel.

> *If there is no change to the message output, then the code in use is insensitive to polarity reversals.*

Note that the LINE-CODE DECODER requires, for successful decoding, an input signal of amplitude near the TIMS ANALOG REFERENCE LEVEL (±2 volt pp). In normal applications this is assured, since it will obtain its input from the DECISION MAKER.

procedure

There are no step-by-step Tasks for you to perform. Instead, it is left to you to ensure that (in the approximate order indicated):

T1. you read the *TIMS Advanced Modules User Manual* for more details of the LINE-CODE ENCODER and LINE-CODE DECODER modules than is included here.

T2. you select a short sequence from the transmitter message source

T3. at least initially you synchronize the oscilloscope to show a snapshot of the transmitter sequence. Later you may be interested in eye patterns ?

T4. examine each code in turn from the encoder, confirming the transformation from TTL is as expected. On the other hand, and far more challenging, is to determine what the law of each transformation is without help from a Textbook or other reference.

T5. of significant interest would be an examination of the power spectra of each of the coded signals. For this you would need data capturing facilities, and software to perform spectral analysis.

T6. and so on ············

resetting

Resetting of the LINE-CODE ENCODER and the LINE-CODE DECODER after the master clock is connected, or after any clock interruption, is strictly not necessary for *all* codes. But it is easier to do it for *all* codes rather than remember for which codes it is essential.

For more details refer to the *TIMS Advanced Modules User Manual*.

Tutorial Questions

Q1 *why introduce the complications of line encoding in a digital transmission system ?*

Q2 *apart from the inevitable delay introduced by the analog filter, did you notice any other delays in the system ? You may need this information when debugging later experiments.*

Q3 *an important function of many line encoders is the elimination of the DC component. When is this desirable ?*

13. ASK - AMPLITUDE SHIFT KEYING

ACHIEVEMENTS: *generation and demodulation of an amplitude shift keyed (ASK) signal.*

PREREQUISITES: *it would be advantageous to have completed some of the experiments in **Volume A1** involving linear modulation and demodulation.*

EXTRA MODULES: *DECISION MAKER*

Preparation

generation

Amplitude shift keying - ASK - in the context of digital communications is a modulation process which imparts to a sinusoid two or more discrete amplitude levels [38]. These are related to the number of levels adopted by the digital message.

For a binary message sequence there are two levels, one of which is typically zero.

Thus the modulated waveform consists of bursts of a sinusoid.

Figure 1 illustrates a binary ASK signal (lower), together with the binary sequence which initiated it (upper). Neither signal has been bandlimited.

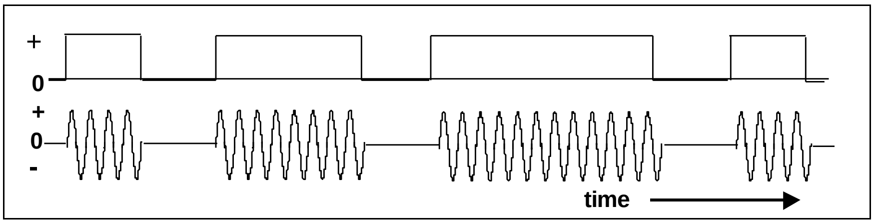


Figure 1: an ASK signal (below) and the message (above)

There are sharp discontinuities shown at the transition points. These result in the signal having an unnecessarily wide bandwidth. Bandlimiting is generally introduced before transmission, in which case these discontinuities would be 'rounded off'. The bandlimiting may be applied to the digital message, or the modulated signal itself.

The data rate is often made a sub-multiple of the carrier frequency. This has been done in the waveform of Figure 1.

[38] also called on-off keying - OOK

One of the disadvantages of ASK, compared with FSK and PSK, for example, is that it has not got a constant envelope. This makes its processing (eg, power amplification) more difficult, since linearity becomes an important factor. However, it does make for ease of demodulation with an envelope detector.

A block diagram of a basic ASK generator is shown in Figure 2. This shows bandlimiting following modulation.

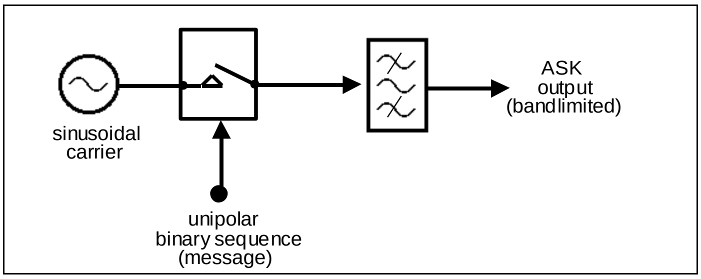


Figure 2: the principle of ASK generation

The switch is opened and closed by the unipolar binary sequence.

bandwidth modification

As already indicated, the sharp discontinuities in the ASK waveform of Figure 1 imply a wide bandwidth. A significant reduction can be accepted before errors at the receiver increase unacceptably. This can be brought about by bandlimiting (pulse shaping) the message *before* modulation, or bandlimiting the ASK signal itself *after* generation.

Both these options are illustrated in Figure 3, which shows one of the generators you will be modelling in this experiment.

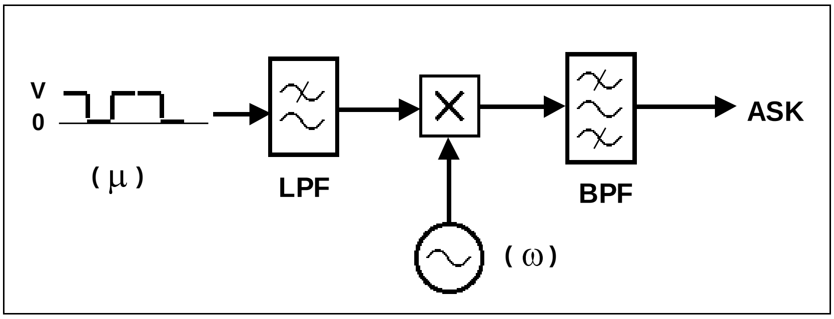


Figure 3: ASK bandlimiting, with a LPF *or* a BPF.

Figure 4 shows the signals present in a model of Figure 3, where the message has been bandlimited. The shape, after bandlimiting, depends naturally enough upon the amplitude and phase characteristics of the bandlimiting filter.

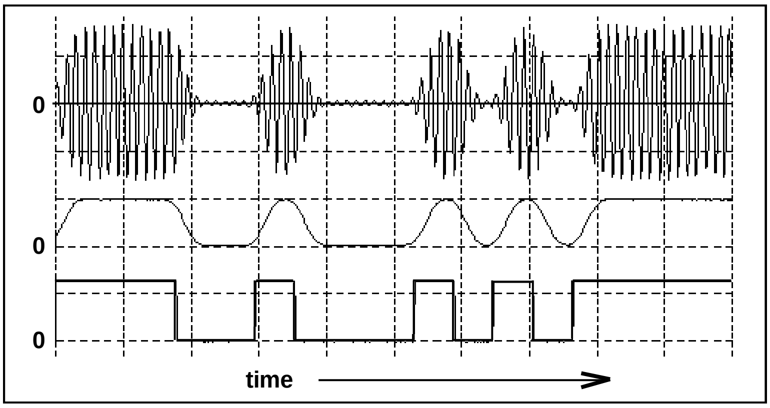


Figure 4: original TTL message (lower), bandlimited message (centre), and ASK (above)

You can approximate these waveforms with a SEQUENCE GENERATOR clocked at about 2 kHz, filter #3 of the BASEBAND CHANNEL FILTERS, and a 10 kHz carrier from a VCO.

demodulation methods

It is apparent from Figures 1 and 4 that the ASK signal has a well defined envelope. Thus it is amenable to demodulation by an envelope detector.

A synchronous demodulator would also be appropriate.

Note that:

- envelope detection circuitry is simple.
- synchronous demodulation requires a phase-locked local carrier and therefore carrier acquisition circuitry.

With bandlimiting of the transmitted ASK neither of these demodulation methods would recover the original binary sequence; instead, their outputs would be a bandlimited version. Thus further processing - by some sort of decision-making circuitry for example - would be necessary.

Thus demodulation is a two-stage process:

1. recovery of the bandlimited bit stream
2. regeneration of the binary bit stream

Figure 5 illustrates.

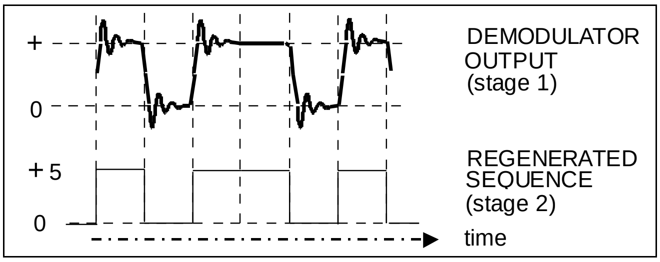


Figure 5: the two stages of the demodulation process

bandwidth estimation

It is easy to estimate the bandwidth of an ASK signal. Refer to the block diagram of Figure 3. This is a DSB transmitter. It is an example of linear modulation. If we know the message bandwidth, then the ASK bandwidth is twice this, centred on the 'carrier' frequency.

Using the analogy of the DSB generator, the binary sequence is the message (bit rate 'μ'), and the sinewave being switched is the carrier ('ω').

Even though you may not have an analytical expression for the bandwidth of a pseudo random binary sequence, you can estimate that it will be of the same order as that of a square, or perhaps a rectangular, wave.

For the special case of a binary sequence of alternate ones and zeros the spectrum will:

- be symmetrical about the frequency of the carrier 'ω'
- have a component at 'ω', because there will be a DC term in the message
- have sidebands spaced at odd multiples of 'μ' either side of the carrier
- have sideband amplitudes which will decrease either side of the carrier (proportional to 1/n, where 'n' is the order of the term).

If you accept the spectrum is symmetrical around the carrier then you can measure its effective bandwidth by passing it through a tuneable lowpass filter. A method is suggested in the experiment below.

You can discuss this when answering Tutorial Question Q2.

Experiment

T1.0 generation

There are many methods of modelling an ASK generator with TIMS. For any of them the binary message sequence is best obtained from a SEQUENCE GENERATOR, clocked at an appropriate speed. Depending upon the generator configuration, either the data bit stream can be bandlimited, or the ASK itself can be bandpass filtered.

Suggestions for modelling the ASK generators are:

T1.1 modelling with a DUAL ANALOG SWITCH

It is possible to model the rather basic generator shown in Figure 2.

The switch can be modelled by one half of a DUAL ANALOG SWITCH module. Being an *analog* switch, the carrier frequency would need to be in the audio range. For example, 15 kHz from a VCO. The TTL output from the SEQUENCE GENERATOR is connected directly to the CONTROL input of the DUAL ANALOG SWITCH. For a synchronous carrier and message use the 8.333 kHz TTL sample clock (filtered by a TUNEABLE LPF) and the 2.083 kHz sinusoidal message from the MASTER SIGNALS module.

If you need the TUNEABLE LPF for bandlimiting of the ASK, use the sinusoidal output from an AUDIO OSCILLATOR as the carrier. For a synchronized message as above, tune the oscillator close to 8.333 kHz, and lock it there with the sample clock connected to its SYNCH input.

This arrangement is shown modelled in Figure 6.

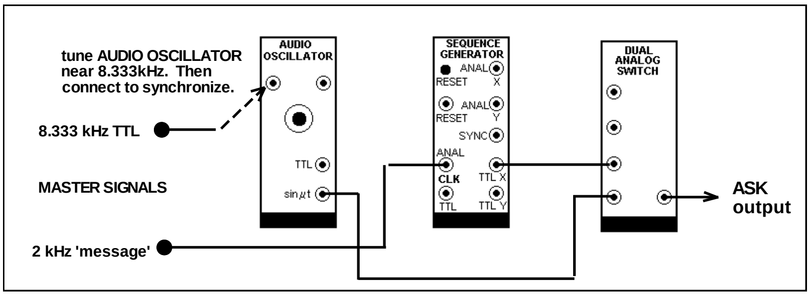


Figure 6: modelling ASK with the arrangement of Figure 2

Bandlimiting can be implemented with a filter at the output of the ANALOG SWITCH.

T1.2 modelling with a MULTIPLIER

A MULTIPLIER module can be used as the switch. The carrier can come from any suitable sinusoidal source. It could be at any available TIMS frequency.

The other input to the MULTIPLIER needs to be the message sequence.

Neither the TTL nor the analog sequence is at an appropriate voltage level. Each requires amplitude scaling. This can be implemented in an ADDER, which will invert the sequence polarity. DC from the VARIABLE DC module can be used to re-set the DC level. The required signal will be at a level of either 0 V or +2 V, the latter being optimum for the (analog) MULTIPLIER.

This arrangement is shown modelled in Figure 7.

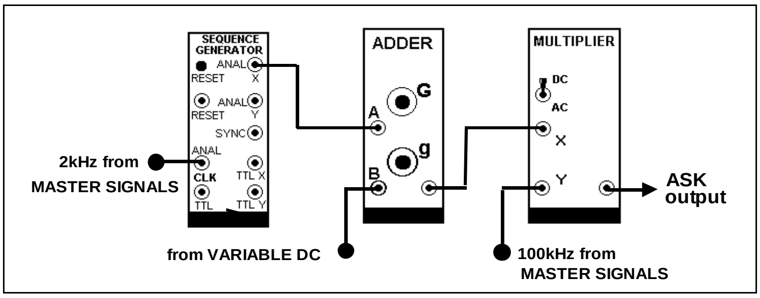


Figure 7: modelling ASK with the arrangement of Figure 3.

The operating frequency of the modulator of Figure 7 is not restricted to audio frequencies. Any carrier frequency available within TIMS may be used, but remember to keep the data rate below that of the carrier frequency.

For a synchronous system (ie, message and carrier rates related, so as to give 'stable' oscilloscope displays):

- clock the SEQUENCE GENERATOR from the 2 kHz message (as shown), or the 8.333 kHz sample clock.
- use a 100 kHz carrier (as shown), or an AUDIO OSCILLATOR locked to the 8.333 kHz sample clock.

Any other combination of data clock and carrier frequency, synchronous or otherwise, is possible (with this model); but not all combinations will generate an ASK signal. Try it !

Bandlimiting can be implemented with a filter at the MULTIPLIER output (a 100 kHz CHANNEL FILTERS module), or the bit sequence itself can be bandlimited (BASEBAND CHANNEL FILTERS module).

T2.0 bandwidth measurement

Having generated an ASK signal, an estimate of its bandwidth can be made using an arrangement such as illustrated in Figure 8. The bandwidth of the lowpass filter is reduced until you consider that the envelope can no longer be identified.

This will indicate the *upper* frequency limit of the signal. Do you think it reasonable to then make a declaration regarding the *lower* frequency limit ?

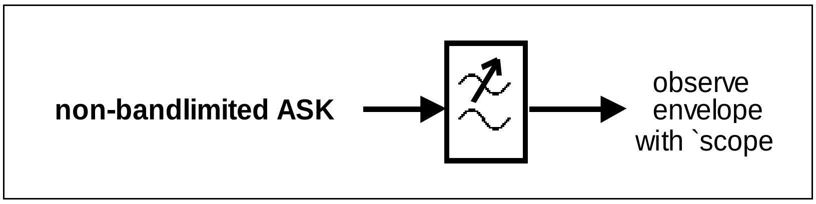


Figure 8: ASK bandwidth estimation

The arrangement of Figure 8 is easy to model with TIMS. Use the TUNEABLE LPF. But remember to select appropriate ASK frequencies.

T3.0 demodulation

Both asynchronous and synchronous demodulation methods are used for the demodulation of ASK signals.

T3.1 envelope demodulation

Having a very definite envelope, an envelope detector can be used as the first step in recovering the original sequence. Further processing can be employed to regenerate the true binary waveform.

Figure 9 is a model for envelope recovery from a baseband FSK signal.

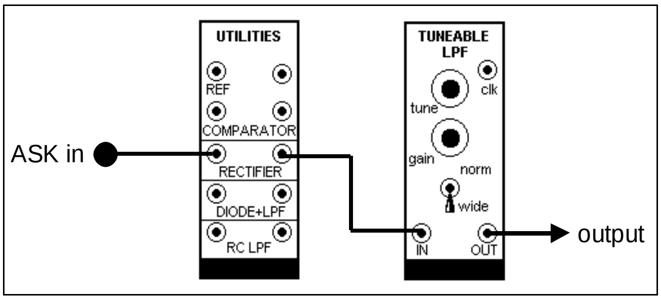


Figure 9: envelope demodulation of baseband ASK

If you choose to evaluate the model of Figure 9, remember there is a relationship between bit rate and the lowpass filter bandwidth. Select your frequencies wisely.

T3.2 synchronous demodulation

A synchronous demodulator can be used for demodulation, as shown in Figure 10. In the laboratory you can use a stolen carrier, as shown.

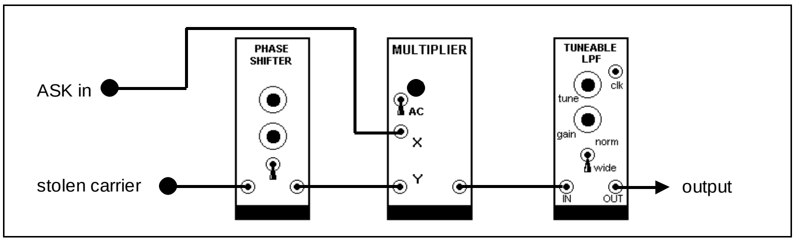


Figure 10: synchronous demodulation of ASK

T3.3 post-demodulation processing

The output from both of the above demodulators will not be a copy of the binary sequence TTL waveform. Bandlimiting will have shaped it, as (for example) illustrated in Figure 4.

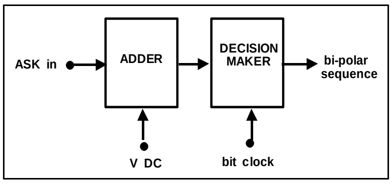


Figure 11

Some sort of decision device is then required to regenerate the original binary sequence. The DECISION MAKER module could be employed, with associated processing, if required. This is illustrated in block diagram form in Figure 11 (opposite).

This model will regenerate a bi-polar sequence from the recovered envelope.

Figure 12 shows the model of the block diagram of Figure 11.

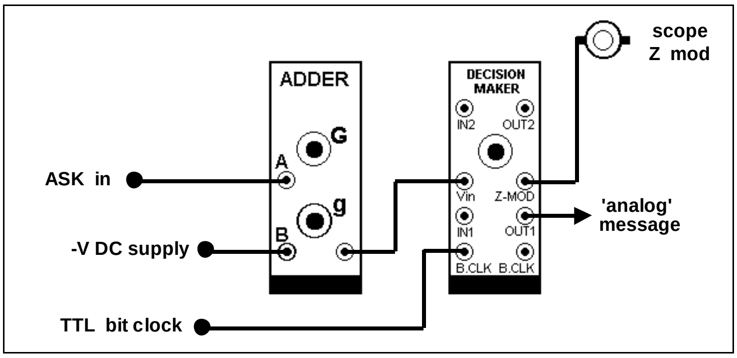


Figure 12: regeneration to a bi-polar sequence

Remember to:

- convert the uni-polar, bandlimited output of the envelope detector to bi-polar (using the ADDER), to suit the DECISION MAKER.
- set the on-board switch SW1, of the DECISION MAKER, to NRZ-L. This configures it to accept bi-polar inputs.
- adjust the decision point of the DECISION MAKER
- in the first instance, use a stolen carrier and bit clock

The output will be the regenerated message waveform. Coming from a YELLOW analog output socket, it is bi-polar ±2 V (not TTL).

The same regenerator can be used to process the output from the synchronous demodulator of Figure 10.

T4.0 carrier acquisition

Rather than using a stolen carrier and bit clock you might like to try recovering these from the received ASK signal.

Tutorial Questions

Q1 *suggest an advantage of making the data rate a sub-multiple of the carrier rate.*

Q2 *discuss your methods of measuring and/or estimating the bandwidth of the ASK signal. Estimate the maximum amount of bandwidth limiting possible, and the trade-offs involved.*

Q3 *the ASK waveform of Figure 1 is 'special' in that:*

a) the bit rate is a sub-multiple of the carrier

b) the phasing of the message ensures that each 'burst' of carrier starts and ends at zero amplitude.

If these special conditions are changed, consider the shape of the waveform at the beginning and end of each burst of carrier. What effect, if any, will this have on the bandwidth of the ASK signal ?

14. BLOCK CODING & DECODING

ACHIEVEMENTS: *viewing of a serial data stream before and after block encoding. Decoding. SNR improvement due to block coding.*

PREREQUISITES: *completion of the experiment entitled* ***PCM encoding*** *in Volume D1.*

ADVANCED MODULES: *PCM ENCODER, BLOCK CODE ENCODER, BLOCK CODE DECODER, LINE-CODE ENCODER.*

Preparation

block coding

This experiment examines the BLOCK CODE ENCODER and BLOCK CODE DECODER modules.

Block coding refers to the technique of adding extra bits to a digital word in order to improve the reliability of transmission. The word consists of the message bits (often called information, or data) *plus* code bits. It may also, as in the present case, contain a frame synchronization bit.

A block code adds bits to existing message bits, or blocks, *independently* of adjacent blocks [39].

In this experiment the blocks will be prepared by the PCM ENCODER module. These blocks were examined in the experiment entitled *PCM encoding*.

PCM encoded data format

When extra code bits are added to a PCM word (initially containing only message bits) then the word will get longer. If the bit rate remained the same then the message bits would arrive at a slower rate than before. To maintain the same message rate the bit rate would need to be increased. This would require an increased transmission bandwidth.

In the TIMS PCM ENCODER module a different scenario has been adopted.

The PCM word has been generated from the input message and placed in a frame of fixed length. These are the *message bits*. Not all slots in the frame are used. When extra *coding bits* are added, they go in the previously unused slots. Thus, in either case (extra code bits or not):

- the frame length remains the same
- the message rate remains the same
- the channel bandwidth will remain the same, as the bit rate has not changed

[39] instead of being distributed over a number of blocks, as, for example, in a convolutional code.

The TIMS arrangement may waste time in the un-block-coded state (there are three unused slots in the frame), and so be called inefficient (which it is). But it is convenient for our purpose.

The PCM ENCODER module was examined in the experiment entitled *PCM encoding* in volume D1.

block code format

The BLOCK CODE ENCODER module is designed to expect input blocks of length eight slots, where some of these slots are empty. These come from the PCM ENCODER module (in the 4-bit mode).

The incoming data frame is illustrated in Figure 1 below.

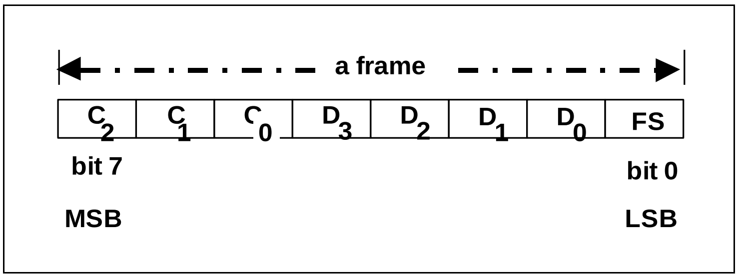


Figure 1: a data frame of eight slots, one per clock period

The message bits are shown as D_3, D_2, D_1, and D_0, where D_3 is the most significant bit of the message.

The frame synchronization bit is shown as FS.

The slots marked C_2, C_1, and C_0 will be used by the BLOCK ENCODER for code bits.

For the BLOCK CODE ENCODER module to function correctly it must always receive three digital signals:

1. TTL binary data in an 8-bit wide frame (typically from a PCM ENCODER in 4-bit mode). The data must occupy frames 4, 3, 2, and 1 (as defined in Figure 1 above).
2. a TTL clock, to which the incoming data is synchronized. Typically this will be at 2.083 kHz (the module is restricted to a clock rate below 8 kHz).
3. a TTL frame synchronization signal FS, which signals the *end* of the frame.

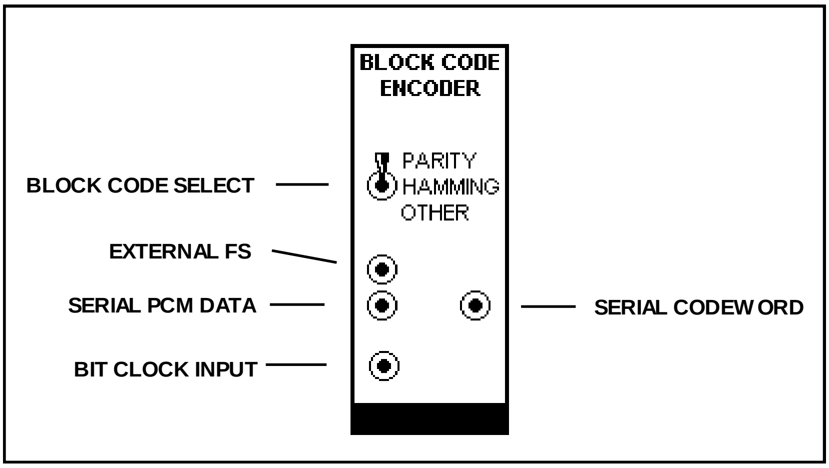


Figure 2: front panel layout - ENCODER

The front panel of the module is illustrated in Figure 2 above. The features should be self explanatory, except for the BLOCK CODE SELECT toggle switch.

block code select

Each BLOCK CODE ENCODER module offers three different coding schemes. These are contained in an EPROM. More than one EPROM is available, any one of which can be installed in the module. The codes they offer are set out in Table 1 below.

EPROM	code 1	code 2	code 3
BLKe1.x	even parity - single bit error detect	Hamming (7,4) - single bit error correct	*Setup - with C_x bit error detect
BLKe2.x	even parity - single bit error detect	Hamming (7,4) - single bit error correct	odd parity - single bit error detect
BLKe3.x	even parity - single bit error detect	Hamming (7,4) - single bit error correct	Cyclic

Table 1: EPROM codes

Any one of the three codes in the installed EPROM can be selected with the front panel toggle switch.

In performing parity checks the FS bit is ignored.

typical usage

In a typical digital communications system, the configuration at the transmitter might appear as in the block diagram of Figure 3 below.

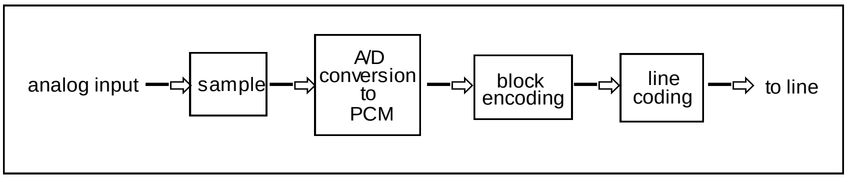


Figure 3: disposition of the block encoder

block decoding

The signals from the BLOCK CODE ENCODER need to be interpreted by a complementary BLOCK CODE DECODER module, the front panel of which is illustrated in Figure 4 below.

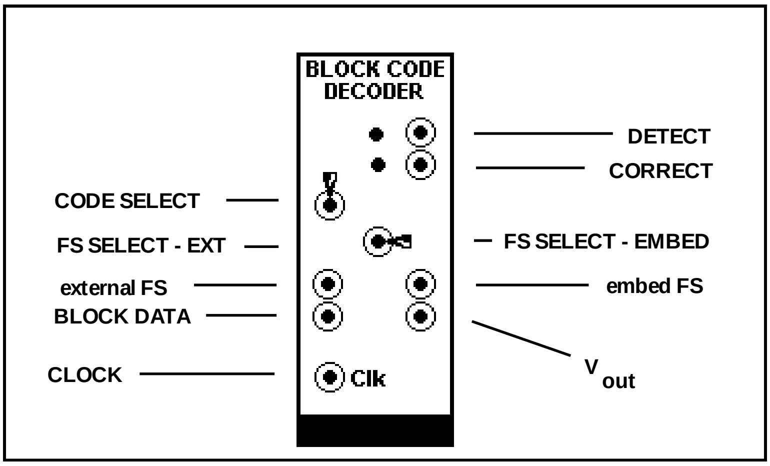


Figure 4: front panel layout - DECODER

The front panel of the decoder module is illustrated in Figure 4 above. The features should be self explanatory, except for the following:

- DETECT: for codes which can detect but not correct errors. The LED flashes when an error is detected, but not corrected. There is a TTL high, one bit wide, at the adjacent socket, during the frame in which the error occurred.
- CORRECT: for codes which can detect and correct errors. The LED flashes when an error is detected and corrected. There is a TTL high, one bit wide, at the adjacent socket, during the frame in which the error occurred.

The DETECT and CORRECT outputs (LED and bit-wide TTL HI) are mutually exclusive.

- FS SELECT - EXT: frame synchronization may be attained by accepting a ‘stolen’ FS signal from the transmitter, patched to the FS input socket.
- FS SELECT - EMBED: frame synchronization may be achieved automatically, using the embedded information in the LSB of the frame itself. For verification the recovered FS signal is available at the FS output socket. When a stolen FS signal is used there is *no output* from this socket.

Experiment

This experiment is intended to help familiarize you with some aspects of the operation of the BLOCK CODE ENCODER. It will also confirm the decoding process performed by the BLOCK CODE DECODER module. It is a necessary preliminary to the experiment entitled *Block coding and coding gain* of this Volume.

encoding

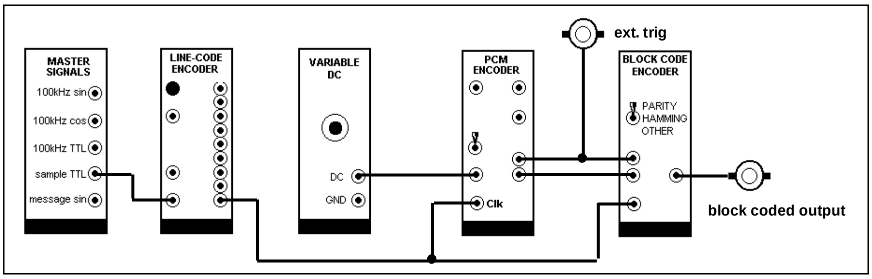


Figure 5: block code encoding

The BLOCK CODE ENCODER requires a TTL clock near 2 kHz. The *TIMS Advanced Modules User Manual* says it must be operated at a clock speed below 8 kHz.

You may have notice that it is customary TIMS practice (but not mandatory) to use a clock locked to the MASTER 100 kHz source. Typically this has been the 8.333 kHz TTL signal from the MASTER SIGNALS module. Since the BLOCK CODE ENCODER requires something lower than this, a convenient source is obtained by dividing this by 4. The LINE-CODE ENCODER module has just such a divider [40] (and typically forms part of a data transmission system). The model of Figure 5 above illustrates this method.

For stable oscilloscope displays from the PCM source a DC message is used, together with a suitable source of external triggering signal.

T1 *patch up the model of Figure 5.*

T2 *set up simultaneous displays of the PCM input, and the block coded output, of the BLOCK CODE ENCODER, over two or three frames. Spend some time investigating different methods of oscilloscope synchronization. Accepting jittering displays is unprofessional ! See Tutorial Question Q2.*

T3 *verify, where possible, that each of the codes has been implemented correctly.*

decoding

Having successfully block encoded a PCM signal, it is time to demonstrate its decoding. For this purpose transmission will be via a direct connection.

[40] the DIGITAL UTILITIES module, and the BIT CLOCK REGEN module also have divide-by-4 sub systems

You will have noticed the ERROR INDICATION front panel LEDs on the decoder. These will be useful when transmission via a noisy, bandlimited channel, with the inclusion of line encoding, is examined in a later experiment. There the benefits of block coding will be demonstrated and evaluated.

Patching for the decoding process is shown in Figure 6.

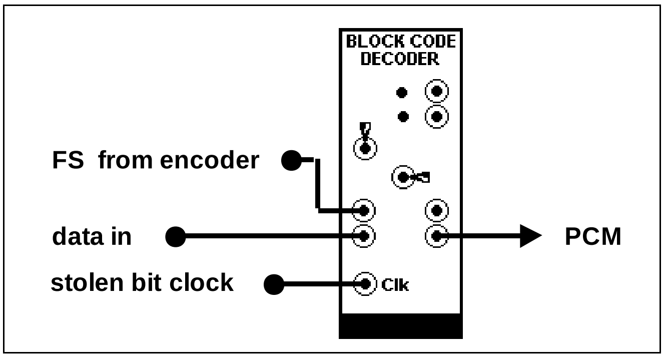


Figure 6: block code decoding

Note that a stolen bit clock is used.

Frame synchronization can be achieved by either a 'stolen' FS signal from the encoder, or by internal decoding of the alternating pattern of 1 - 0 - 1 - 0 - 1 embedded as the LSB of the PCM code word (in location '0' of the frame). This scheme was introduced in the experiment entitled *PCM decoding*.

T4 *patch up the BLOCK CODE DECODER according to Figure 6. This uses the 'stolen' frame synchronization signal FS from the transmitter, connected to the* EXT. FS *input, and selected with the front panel toggle switch* FS SELECT*.*

T5 *verify that successful decoding back to the original PCM is possible for all codes.*

T6 *switch the front panel toggle switch* FS SELECT *to* EMBED*. Confirm that the internal circuitry for extracting the frame synchronization signal FS from the PCM signal itself is operating correctly. Refer to the Appendix to this experiment for more information.*

conclusion

You are now in a position to include block coding in a more complex transmission system (noisy, bandlimited) and to demonstrate its effectiveness in improving the bit error rate. This is the subject of the experiment entitled *Block coding and coding gain*.

During this experiment you should have developed techniques for obtaining oscilloscope displays which show you what you want, without need to constantly adjust and re-adjust the oscilloscope controls. Choice of the appropriate trigger signal for each display is important.

Although, for a DC message, each 4-bit word and added code bits are the same, the alternating pattern of 0-1-0-1- for the FS signal make alternate frames *different*. It would be preferred if the synchronisation technique adopted would always put the same frame first, no matter what the sweep speed.

This is a simple matter to implement. See Tutorial Question Q2.

Tutorial Questions

Q1 *when adding check bits for parity checking, the bits of the alternating pattern 1 - 0 - 1 - 0 - for frame synchronization in the LSB position were ignored. Explain.*

Q2 *explain how dividing the frame synchronization signal by two is often a help in obtaining and maintaining stable, and repeatable, oscilloscope displays in the context of this experiment.*

Appendix

automatic frame synchronization

The BLOCK CODE DECODER module has built-in circuitry for locating the position of each frame in the serial data stream. The circuitry looks for the embedded and alternating '0' and '1' every 8 bits (which occur in the LSB position of each frame).

The search is made by examining a section of data whose length is a multiple of eight bits.

The length of this section can be changed by the on-board switch SW3. Under noisy conditions it is advantageous to use longer lengths.

The switch settings are listed in Table A-1 below.

left toggle	right toggle	groups of eight bits
UP	UP	4
UP	DOWN	8
DOWN	UP	16
DOWN	DOWN	32

Table A-1: synchronization search length options