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TITLE

'THE DESIGN OF AN ELECTRONIC MUSIC SYNTHESIZER'

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The Design Basis of an Electronic Music Synthesizer

SUMMARY

Initially this thesis examines the basic design considerations of electronic music synthesizers. It is pointed out that, irrespective of how sophisticated the musical instrument, it is necessary for the operator to select sounds by means of physical movements. If one of a large number of sounds is to be selected, then one of that large number of physical operations must be selected by the operator of the musical instrument. The implications of using taperecording techniques is considered. It is concluded that, at the outset, a musical instrument should be designed as either a monitored instrument where the sounds are heard as they are being selected, or as a programmed instrument where the instrument is adjusted at leisure but without the benefit of hearing the corresponding sound at the same time as the instrument is adjusted.

The thesis gives a design of a monitored music synthesizer. The design of each unit is related to the basic design considerations. The question of how to interconnect the units is examined, as is the relative importance of the sound sources and sound manipulators. The designs of the individual units are very different, so that different sounds are available to the musician. The four sound sources use the time domain, frequency domain, digital division and white noise as bases for the design approach. The sound manipulators produce lower frequency components as well as the more usual higher frequency components, and a voltage controlled filter is described.

The control units use a variety of sources of control; a hand operated lever with three degrees of freedom, a keyboard where combinations of notes as well as pressure are used, a foot controller, and even the sound of the voice is used. Two designs are given for reference oscillators which can be used when adjusting the keyboard controllers.

Each of these units is designed so as to make the greatest number of sounds available to the musician.

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Mr. P. Tahourdin gave considerable assistance when developing the engineering basis for the design of the synthesizer. This included many discussiond on the musician's requirements for a synthesizer, as well as access to the University of Adelaide's Moog synthesizer. Mr. D. C. Pawsey gave technical supervision which included many helpful suggestions and critical comments. The thesis owes a great deal to this guidance. Dr. K. Kikkert continued this supervision for one year in Mr. Pawsey's absence.

DECLARATION

This thesis contains no material which has been accepted for the award of any other degree or diploma in any university and that, tb the best of the candidate's knowledge and belief, the thesis contains no material previously published or written by another person, except where due reference is made in the text of the thesis.

CHAPTER 1

1. Design Basis for the Music Synthesizer.

1.1 An Engineering Approach to the Design of an Electronic Music Synthesizer

1.1.1 Background of Development [1]

The role of an engineer in developing an electronic music synthesizer can best be seen in its historical context. Before the technology of steel production was developed a large range of musical instruments were not available to musicians. The development of this technology allowed artisans to provide instruments which are commonplace today. The initial acceptance of the instruments involved discussions which are familiar to the electronic musician of today; the music from pianos was not really music.

The full musical implications of the development of electronic technology are considerably greater than those of previous technical innovations. There is extremely rapid growth in this field, and it has direct and convenient application to the generation of musical sounds. The impact of electronic transmission and storage of music can be viewed in this context. Even the reproduction of sound involves aesthetic considerations. The sound actually comes from a loadspeaker and can be processed by tone controls. Electronic musical instruments involve more radical aesthetic considerations. This may mean the realization of the full potential of electronic techniques may be a gradual process.

The extensive advantages of electronic techniques can obscure the inherent limitations. The objective of this first chapter is to outline these limitations, and to discuss their implications for the design of an electronic instrument.

1.1.2 Limitations of Conventional Musical Instruments

Performers using conventional musical instruments have to be extremely skilful. The training required to aquire this skill means that some musicians dedicate their working lives to performing works which are written by other people. Often the composer must rely upon a performer to realize his work. This means interpretation by the performer is inevitable, and the composer may prefer this is avoided. There is an inherent conservatism in this situation; the composer is limited by the range of skills and instruments available to realize his work, and the means of communication between the composer and the performer limits the degree to which the musical work can be specified precisely. A conventional musical score specifies only 12 frequencies to the octave, has only limited notation to give sound intensities and duration, and is completely dependent upon the instrument for the harmonic and transient components of the notes. It is understandable that a composer who contemplates radically new sounds would turn to electronic techniques.

1.1.3 Selection of a sound using any musical instrument

Although electronic techniques allow considerable scope for development they do not change some basic engineering requirements for musical instruments. All musical instruments are altered physically by an operator to produce different sounds. If one of two distinct sounds is required, then one of at least two different physical operations must be performed. More generally, if it is desired to select a particular sound from N distinct sounds, then it is necessary to perform one of at least N distinct physical operations. This fact applies to all musical instruments, electronic or conventional. It is independent of the complexity of the sound produced by the instrument and is independent of the sophistication of the instrument. It applies of the sound is selected at reduced speed using tape recording techniques, or if a number of tracks are taped independently and then superimposed.

The musician would like to be able to choose one of a large number of sounds. He would like to be able to do this without having to aquire an unnecessarily high degree of performing skill. However, if the sound is selected from a large number of sounds, there must be one of at least that large number of physical operations performed. This would seem to necessitate a degree of performing skill.

1.1.3.1 Reducing the recording speed during performance[1]

Musicians have used tape recording techniques in an attempt to minimize the requirement for performing skill.[1] If the rate at which the sound is recorded is slowed down, then the rate at which the operations are performed is slowed down also. However, one important difficulty does arise. Although this technique gives the operator more time to select from the same number of physical operations, the sound which will appear in the final work is no longer heard as it is being selected; it is no longer monitored.

1.1.3.2 Multichannel recording[2]

A similar situation arises with multichannel recording. If the number of channels is small it may be possible for the musician to judge how a given channel will sound when replayed with the others. This may provide some useful extension of the time to perform the physical operations. However it would be necessary to use a very large number of channels of this technique were to compensate for a large degree of performing skill. If a large number of channels is used it would be impossible to judge how a given channel would sound when it was replayed with the others. Therefore the final sound is not monitored in this case either.

1.1.4 Monitored musical instruments

Monitoring the sound produced by a musical instrument as it is selected is a very important aspect of selecting the sound. To illustrate this, consider the difficulty of playing a piano if the sound is heard one minute after the corresponding note is depressed. A monitored instrument has the advantage that the operator has the sound available to provide feedback when selecting a sound. It also has the disadvantage that the sounds must be selected at a rate which is similar to the rate the appear in the final work. On the other hand, if the sound is not heard as it is being selected the delay between the physical operation and hearing the sound makes the selection of a particular sound much more difficult.

1.1.5 Comparison between monitored and programmed musical instruments

There is a distinction between monitored and programmed musical instruments which is useful when making design decisions. With a monitored musical instrument the sound which results from a physical operation is heard at the same time as the operation is performed, whereas the final sound from a programmed instrument is heard some time after the physical operation selects that sound.[3] This distinction is reflected in the different emphases in the design criteria of these instruments. With a monitored instrument it is important to use as many different types of movement of the operator as possible. The foot pedals of a large organ provide extra control over the sound produced by the instrument, and this could not be obtained using only the hands. The piano-forte keyboard responds to the force with which the key is depressed as well as to which key is selected. This is an important advantage of the piano-forte over the harpsichord. If this extra differentiation is achieved it becomes possible to select between a larger number of physical operations, and therefore between a larger number of sounds.

With a programmed musical instrument is it important that the operator can look at the controls and estimate the sound which will be produced. In this way it is possible for the operator to obtain a particular sound with a minimum number of attempts or iterations.

There are two design criteria which are fundamental to the design of both monitored and programmed musical instruments. These are examined in section 1.2. The design implications of these criteria for the monitored instruments are then discussed in more detail in section 1.3, and then the design is outlined in section 1.5. The design implications for programmed musical instruments are discussed in section 1.4 where a design is proposed. The monitored synthesizer designed for this thesis is compared with a commercial synthesizer in section 1.6. Emphasis is given to the application of the design criteria.

1.2 Two Design Criteria for Musical Instruments

The two design criteria are;

i. A large number of soundsshould be available to the musician,

and ii. Selecting from these sounds should be as convenient for the musician as possible.

The musician should be able to choose one of a large number of sounds with a minimum of training. Some of the aesthetic considerations of these criteria are discussed. The provision of a large number of sounds has two implications.

1. The instrument should be capable of producing a wide range of types of sound.

Usually the musician does not want the scope of the available sounds limited by the instrument. However there is an alternative point of view. If the musicians are forced to work within limitations, then the available sounds are explored in considerable detail. Our musical heritage provides an excellent example of a restriction of the types of sound used. For over two centuries the even tempered 12 tone scale has been used almost exclusively. Its use has been considerably refined. This development would not be as intense if other frequency intervals were commonly used as well. This means many subtle combinations of the 12 tone notes would never have been utilized if other frequency intervals were also available.

On the other hand, the engineer is in no position to design for restrictions in the type of sound available. He is in no position to decide which restrictions are likely to be worthwhile because that is a question of aesthetics. This means the discipline to study a particular type of sound intensively must come from the musician.

2. The second implication is that the musician should be able to select a sound with a great degree of resolution or subtlety.

Consider again the use of pressure variations on the keys of a keyboard. With the conventional piano-forte it is possible to control a wide range of amplitudes using a range of pressures. It is simple to extend this using electronic techniques. The pressure can be used alone, or it can be combined with its rate of attack or decay. The combination can control a number of sound parameters such as amplitude, frequency or harmonic content of a note.

However it is a question of aesthetics whether it is desirable to have a wide range of envelopes for the notes. It might be desired to have the envelopes identical for aesthetic reasons. In principle this could be obtained by depressing the keys with identical rates and pressures. With a piano-forte this is the only way to obtain identical envelope shapes, because the relationship between the key pressure and the sound is fixed. However with electronic instruments this relationship can be altered. It is possible to adjust the pressure control to produce little or no effect on the sound. In this way the musician who wants notes with identical amplitude, for example, can obtain these with a minimum of training or difficulty. However, the design compromises of the instrument have been toward providing the option of control, and to this extent the musician who always wants regular amplitudes has not been catered for.

The second design criterion was that selecting from these sounds should be as convenient for the musician as possible. Again this is the most likely request from the musician, and again there is an alternative point of view.

The convenience of selecting a particular type of sound involves minimizing the difficulty of manipulating the controls. It also involves producing a predictable relationship between the position of the controls and the sound which is produced.

Part of the listener's musical experience may include appreciating the skill of the operator. For example there is considerable difficulty blowing many wind instruments. After a particularly skilful performance the listener may marvel ther there were no errors, and this may be a significant part of his total musical experience. However, it would seem the significance of skill would always be present, even if it were put to more useful effect by differentiating between sounds. Alternative skills could always be provided in any case.

Another aspect ot the convenience of the selection of a sound is its predictability. If the type of sound produced by the instrument changes unpredictably it becomes much more difficult or even impossible to obtain a particular sound. On the other hand in this case the musician would find a variety of types of sound which he may not otherwise imagine. This would expand his library of available sounds and would therefore help composition.

However, if the musician then wants to use the musical instrument to alter that sound slightly, or to produce a given sound, it becomes essential to minimize this element of surprise. To some extent the musician could compensate for this predictability by random experimentation.

The point of view in common with both of these design criteria is that the musician should be able to choose one of a large number of sounds with a minimum of training. If, for any reason, he would like the type of sound restricted, or the difficulty of choosing a sound increased, or fine variations in a sound parameter such as amplitude avoided, then this decision involves aesthetics and is properly his. A musical instrument could be specifically designed to meet any or all of these particular requirements, but the criteria chosen seem the most suitable and most challenging. Nevertheless, it is important to have the assumptions which involve aesthetics clearly identified.

1.3 Design of a Monitored Music Synthesizer.

With a monitored musical instrument the sound in the final work is heard as it is being selected. It follows that the physical operations which are necessary to select the sound must be performed at the same time as the sound is heard. However the number of sounds which can be discerned by the ear in a given time interval greatly exceeds the number of distinct physical operations which can be performed during that time. In other words the rate at which the ear can differentiate between sounds greatly exceeds the rate at which it is possible to select distinct physical operations. Therefore only a small proportion of all possible sounds will ever be available from a monitored musical instrument whilst one operator is using it. The particular sounds which are available from the monitored instrument can be changed by readjustment of the instrument.

Most musical instruments are programmed to some extent because some physical operations on monitored instruments do not produce sound in real time. One example is the tuning of violin strings. The frequency of each string is adjusted before the performance, and this makes a certain group of sounds available to the musician. In some established works the music is stopped and the violin strings retuned to give different frequencies and therefore a different group of sounds.

In some cases there is no rigid distinction between a monitored and a programmed musical instrument. The stops of an organ illustrate the programming of a monitored instrument. These stops can be altered at the beginning of a section of music to provide particular types of sound. The stops can also be altered while the music is being played, and this shows there can be no precise definition of the difference between the setting up of a musical instrument and its real time performance. However there is usually a considerable difference between the two operations as there is between the tuning of and the bowing of a violin. This means a useful distinction can be made, and this is reflected in the differing design approaches to the two instruments.

There is in fact a graduation between a programmed instrument and a monitored instrument. After the programmed instrument is set up the operation of pressing a button to hear the sound as the same as those which characterize a monitored instrument. Similarly, setting up a monitored instrument is the same type of operation which characterizes a programmed instrument. In practice there is a very useful difference between the two types of instruments.

The limited rate of selection of sounds by physical operations means that a small proportion of the sounds which can be distinguished by the ear are available with any one setting or arrangement of the monitored instrument. However, since the monitored instrument can be set up to make different sounds available, in principle anyway, it would be possible to produce any sound at all from a monitored instrument. This instrument would have to be repeatedly set up to obtain these sounds.

It follows there are two design implications for monitored instruments when maximizing the number of sounds available to the musician. The physical operations which select the sounds should be designed to enable the operator to distinguish between a large number of sounds, and the instruments should be designed so that the range of sounds available should be as large as possible.

If sections of electronic equipment are designed to be interconnected in many different ways, the number of sounds available from a given amount of electronic equipment is greatly extended. This technique was introduced by R. A. Moog and others.[4] An example of the advantage when trying to produce a large range of sounds with a given amount of electrical equipment occurs with a filter. This filter could follow an oscillator and alter its harmonic content. It could also follow a white noise generator and produce completely different sounds, resembling sounds from the sea. Both these sounds are available without additional electrical equipment.

Interconnection of units means that the signal levels must be standardized. There is also the aspect of control of these units. The frequency of an oscillator could be controlled by a keyboard, or by the signal from another oscillator, or a combination of both.

This technique can be used to alter the harmonic content of the oscillator because of the effects of frequency modulation. Therefore the ability to interchange the means of control greatly increases the number of sounds available. This means all the controlled functions must depend upon voltages which, ofcourse, must be compatible with the signal levels. The use of voltage controlled circuits is not as convenient for the designer as direct forms of control. It is much simpler to use a potentiometer to control the amplitude of a signal than it is to use a voltage controlled attenuator circuit. However if an attenuator circuit is used it becomes possible to use a voltage from pressure on a keyboard or pedal or some combination. The method of control becomes much more flexible and the number of sounds available from a given amount of equipment is greatly increased.[5]

This design approach contrasts sharply with expensive and intricate electronic syntheses of sounds which are produced by existing instruments. It is possible to simulate a note from a piano using electronics. However this is very difficult. Second order terms in the mounting of piano strings introduce a slight frequency shift in the harmonics. Extensive circuitry is necessary to attempt the synthesis of the sound from a piano, and this equipment could be used to provide a large number of sounds which are characteristic of electronic techniques rather than strings. The use of characteristically electronic sounds is an important aspect of an art form which uses an electronic medium. For the same reason the circuits should not be elaborate just to produce sounds which are very different from familiar sounds.

In order to provide a great range of sounds with a given amount of equipment, the functions of the units which have been designed are very different from each other. There has been a systematic attempt to use a range of electronic techniques when producing and manipulating the sounds.

The other aspect ot making many sounds available to the musician is to enable the selection of a particular type of sound with a great degree of resolution or subtlety. The control units allow the physical operation to differentiate between the sounds. As many aspects of these physical operations should be used as possible; foot controllers, pressure on keyboards, lever controllers with three degrees of freedom. Again this design approach can be contrasted with the use of triggering circuits which provide envelope shaping. The inevitable restriction of this technique of only having the trigger signal to indicate the musician's intention becomes apparent when all the envelope shapes are exactly uniform, until a time consuming readjustment is made to the controls.

The second design criterion, to make the sounds available to the musician with a minimum of training, means the control movements should be conveniently related to the sound; position logarithmically related to frequency and amplitude. Also the operator should find the minimum difficulty in appreciating the functions of each of the units. For instance the result of using a filter would be predictable, whether is was used with an oscillator or with a white noise source. It is relatively simple to design circuits which produce sounds which cannot be altered predictably. However the design must ensure the operator can predict which sounds are available and learn how to adjust the instrument to get them. In this respect the design of the monitored instrument is the same as the design of the programmed instrument. The difference is that there is no systematic attempt to produce any sound at all from the monitored instrument, rather to produce a large number of sounds which are convenient to generate electrically and which can be obtained and controlled by an operator.

1.4 Design Outline of a Programmed Instrument

A programmed musical instrument is one where the sound in the final work is not heard as the instrument is adjusted. Only after some considerable delay is the sound heard, and therefore only after this delay can the sound be used to make changes. In this case it is possible to aim at producing any discernable sound because the instrument is not limited by the rate at which the operator can distinguish between physical operations.

The design would take advantage of the fact that the sound is produced for an ear rather than an oscilloscope. Although the bandwidth of the ear is 16 kHz, it is insensitive to phase relationships at normal amplitudes because its operation relies on resonance effects. Therefore it is more suitable to specify frequency power densities rather than waveforms. The function giving frequency power density would be chosen to suit the characteristics of the ear.

In order to make the sounds convenient for the musician to obtain it would be important that the musician could predict the sound which will be produced by simply looking at the instrument. A visual display would be an important part of the instrument. The failure to provide a visual display is a major shortcoming of modern programmed musical instruments using computers. This visual display would form an integral part of the method of storing and generating the sound.

The instrument could use a small computer or microprocessor in conjunction with a light scanner. Each setting of the instrument would involve producing a two dimensional display of frequency power density as a function of time. The intensity of shading would give the power density, and its position would give the frequency and the time at which it occured. This display could be several meters long and one meter wide. This would give several seconds of sound. The scanning interval of time would be chosen to suit the resolution of the ear, and the small computer or microprocessor would perform the fast fourier transform on the frequency power density. Any amount of computing time would be available, and each segment could be stored on tape or stored digitally.

Each segment of sound could be readjusted by the musician with the aid of the visual display. Photographic techniques could be used to record the sound as well as give the musical score. The sound which is stored in this way could be altered after it was replayed.

This design approach to the programmed musical instrument can be contrasted with the use of specialized programming languages on large computers.[6]

1.5 Outline of the Monitored Instrument Designed.

It is convenient to divide the units of a music synthesizer into three classes.

1. Sound Sources

These units are characterized by having signal output only. They also have control voltage inputs. The units are:

- 3.2 Variable Harmonic Oscillator
- 3.3 Variable Waveform Oscillator
- 3.4 Digital Division Oscillator
- 3.5 White Noise Source

2. Sound Manipulators

These units are characterized by having signal inputs as well as signal outputs. They also have control inputs. The units are:

- 4.3 Voltage Controlled Filter
- 4.4 Multiplier
- 4.5 Harmonic Generator
- 4.6 Reverberation Unit
- 4.7 Position Control
- 4.8 Frequency Divider

3. Control Units

These units have control voltage outputs and do not have signal outputs. The Resistance Board Controller and the Sound to Voltage Unit can have signal inputs, and the Control Voltage Manipulator has a control voltage input.

The control units are:

5.2 Conventional Keyboard

5.3 Binary Keyboard

5.4 Lever Controller

5.5 Foot Controller

5.6 Resistance Board Controller

5.7 Sound to Voltage Unit

5.8 Control Voltage Manipulator

In addition to these units the synthesizer needs power output amplifiers and a power supply. There are also two frequency standard oscillators. There is considerable attention given to the method of interconnection of the units. This forms a major part of the cost and this leads to a compromise between cost and complexity. This peripheral equipment is discussed in chapters 2 and 6.

For the remainder of this chapter there is a comparison with the designed synthesizer with a commercially available synthesizer.

1.6 Comparisons Between the Designed Music Synthesizer and the Moog synthesizer.

The first design criterion is to maximize the operator's control of the sound produced. The Binary Keyboard built for this thesis (5.3) produces a voltage which depends upon the finger pressure of a key. An independent voltage output depends upon which keys are depressed. The control voltage which is related to pressure is available to alter the amplitude of a note. There is no pressure control voltage from the keyboard of the Moog synthesizer. Nevertheless it is still possible to control the envelope of a note. A trigger signal is produced by the Moog keyboard, and this starts the operation of an envelope shaper. This shaper can be adjusted to alter the attack rate, the length of time which the note is sustained, and also the rate of decay of the envelope of the note. Even more complex sequences are available by cascading these envelope shapers. The envelope shaper is actuated whenever a key is depressed. However the musician is still severely restricted. Each note must have the same envelope shape. This applies until the envelope shaping unit is readjusted, and this adjustment takes a considerable amount of time. Therefore, although it is possible to control the shape of the envelope of a note on the Moog, it is important not to overlook the fact that the envelope shape cannot be altered without a time consuming readjustment of the synthesizer. This restriction follows inevitably from the limitations of the control voltages which are provided by the keyboard. An analog control voltage can differentiate between more sounds than a trigger pulse. In the Binary Keyboard the force on the keyboard was used to give an analog voltage rather than an impulse.

Another design criterion is to maximize the types of sound available. The Moog synthesizer has nine identical oscillators and a tenth which is only slightly different. The proposed synthesizer has only three oscillators which each use a completly different design approach. The Variable Harmonic Oscillator (3.2) is based on providing continuous control of constituent waveforms which have specific harmonic contents. The Variable Waveform Oscillator (3.3), enables the operator to control the waveform of the generated signal directly, and the Digital Division Oscillator (3.4) provides combinations of signals which are obtained by dividing a given high frequency.

Each oscillator generates two signals of the same frequency but with differing harmonic contents. It is possible to change from one to the other using voltage control. In this way the harmonic content can be controlled by the pressure on a keyboard.

There is a similar diversity in the type of sound manipulators provided with the proposed design. The Harmonic Generator (4.5) generates harmonics which depend upon the amplitude of the input signal. The Frequency Divider (4.8) generates submultiple frequencies of an input frequency. The resources of the Moog synthesizer are allocated to a programable sequencer which is set up to provide a finite number of control voltage levels which are sequenced. Every time this unit is changed there must be a careful and lengthy adjustment of the unit, and this process can be contrasted with the way the proposed synthesizer would be used.

NOTES FOR CHAPTER 1

1. See reference 11 (pp 14 et seq) for a discussion of the historical development of electronic musical instruments

2. Reference 8 gives an account of the role of tape-recording techniques in an electronic music studio. It discusses the use of groups of sounds (multichannel recording) as well as varying the playback speed of the tape.

3. The classic programmed musical instrument, the RCA synthesizer at Columbia University, is described in reference 20. The use of large digital computers is discussed in outline in reference 10 (pp 359 et seq.). Reference 7 describes a special computer program (the MUSIC4BF) for a programmed musical instrument.

4. The voltage control technique is described by R.A. Moog in reference 17.

5. The role of the feedback of sound information to the musician is mentioned in reference 18 which is a very recent article. (1977)

6. Reference 10 contains a discussion of the role of the computer in the generation of music using random notes. Reference 11 gives a detailed description of a computer language.

CHAPTER 2 GENERAL DESCRIPTION OF THE SYNTHESIZER DESIGN

2.1 Introduction

The objective of an electronic synthesizer is to make available a wide range of sounds to the musician; sounds which can be obtained with a minimum of training. The types of sounds should vary widely. It follows the units providing the sounds should use radically different design approaches.

In order to maximize the types of sounds available from a given amount of electrical equiplemt, the equipment is subdivided into units which can be interconnected in many different ways. With this arrangement some of the units are sound sources and others are sound manipulators (with signal inputs as well as signal outputs). The control function must be interchangeable also; voltage levels are used to control the sound generation and manipulation so that the control units can also be interchanged.

Once set up the instrument should be controlled by the musician in real time to select between a large number of sounds. The control units relate the physical movement to the control voltage, and should utilize as many aspects of the movement of the operator. This movement should be conveniently related to the corresponding sound which is produced.

Experience with the instrument should soon enable the musician to infer its capabilities and limitations, and to obtain the desired sound. This means tha operation of the units should be predictable.

The implications of these design requirements are outlined.

2.1.1 Radically Different Design Approaches

Each of the units uses a fundamentally different approach to the generation or alteration of the sound. The sound sources have periodic and nonperiodic waveforms. Of the periodic waveforms one oscillator provides waveforms of different harmonic contents which can be combined. A second oscillator is set up using the time domain waveform. The third oscillator uses submultiples of a given frequency. This signal can have a period in excess of one second when the origional square wave is 1 kHz. Each of these sound sources is convenient electrically when compared with the synthesis of the sound from a piano or other conventional musical instrument. Many types of sounds are provided using limited resources, and these are characteristic of the electronic medium.

Some sound manipulators are linear and others nonlinear. The Filter, Position Control and Reverberation Unit have distinct functions which are all linear. The Harmonic Generator and Frequency Divider are nonlinear and add higher and lower frequency components respectively. The Signal Multiplier gives a very useful acoustic effect. The distinctive electrical functions of these units allows the instrument to provide distinctive sounds.

2.1.2 Control of the sound

The control units enable the musician to obtain the desired sound. The Binary Keyboard uses pressure information. Instead of duplicating the detecting mechanism for every available note, the design recognizes that electronic instruments do not lend themselves to the production of more than one note at a time. An electronic organ requires an extensive duplication of equipment to achieve a result which is rather more characteristic of a mechanical system. The consequence of providing only one note from one oscillator is that only one control voltage can be used to determine frequency. This means that combinations of keys can be used. In this case 10 keys can provide $2^{10} - 1$ notes. If only 10 keys are used the pressure information from the keyboard can be achieved quite conveniently.

The control units include lever controllers with three degrees of freedom and pressure information, Foot Controller to use other limbs, and a Sound to Voltage unit. With the Sound to Voltage unit, two control voltages are produced by the frequency and by the amplitude of an incoming signal. In this way the training of the voice can be used, as well as the scope of using any other sound.

The resistance board controller provides a control voltage which depends upon signals. The relationship to these signals is controlled by the position of a metallic detector on resistance paper.

The control voltage manipulator filters the control voltage, providing derivative and integral components. This provides greater flexibility in the relationship between the sound and the physical movement selecting this sound.

The design of controlling mechanisms especially for electronic musical instruments, with regard for their particular requirements and flexibility, can provide considerably more control than mere imitations of established methods of controlling sound. In many cases the principle is not new; trumpets produce one note at a time and, in part, use combinations of three key positions to control its frequency. The control voltage manipulator is similar in function to the loud and the soft pedals of a piano, although its versatility is much greater. However in the cases of the Sound to Voltage unit, the Lever Controller and the Resistance Board Controller, the flexibility of electronic techniques has provided a radically new approach.

2.2 List of Units in the Synthesizer Design

Section	Unit	Signal		Control	
3.2	Variable Harmonic	0	2	2	0
3.3	Variable Waveform	0	2	2	0
3.4	Digital Division Oscillator	0	2	2	0
3.5	White Noise Generator	0	2	1	0
3.6	Input Amplifier	0	3	0	0
4.3	Voltage Controlled Filter	1	3	2	0
4.4	Multiplier	5	3	0	0
4.5	Harmonic Generator	1	1	1	0
4.6	Reverberation Unit	1	1	0	0
4.7	Position Control	4	-	2	0
5.2	Conventional Keyboard	0	0	0	1
5.3	Binary Keyboard	0	0	0	2
5.4	Lever Controller	0	0	0	3
5.5	Foot Controller	0	0	0	2
5.6	Sound to Voltage Unit	0	0	0	2
5.7	Resistance Board Controller	0	0	0	2
5.8	Control Voltage Manipulator	0	0	4	4
6.2	Analog Frequency Standard	connected directly to			
6.3	Hybrid Frequency Standard	output amplifiers.			

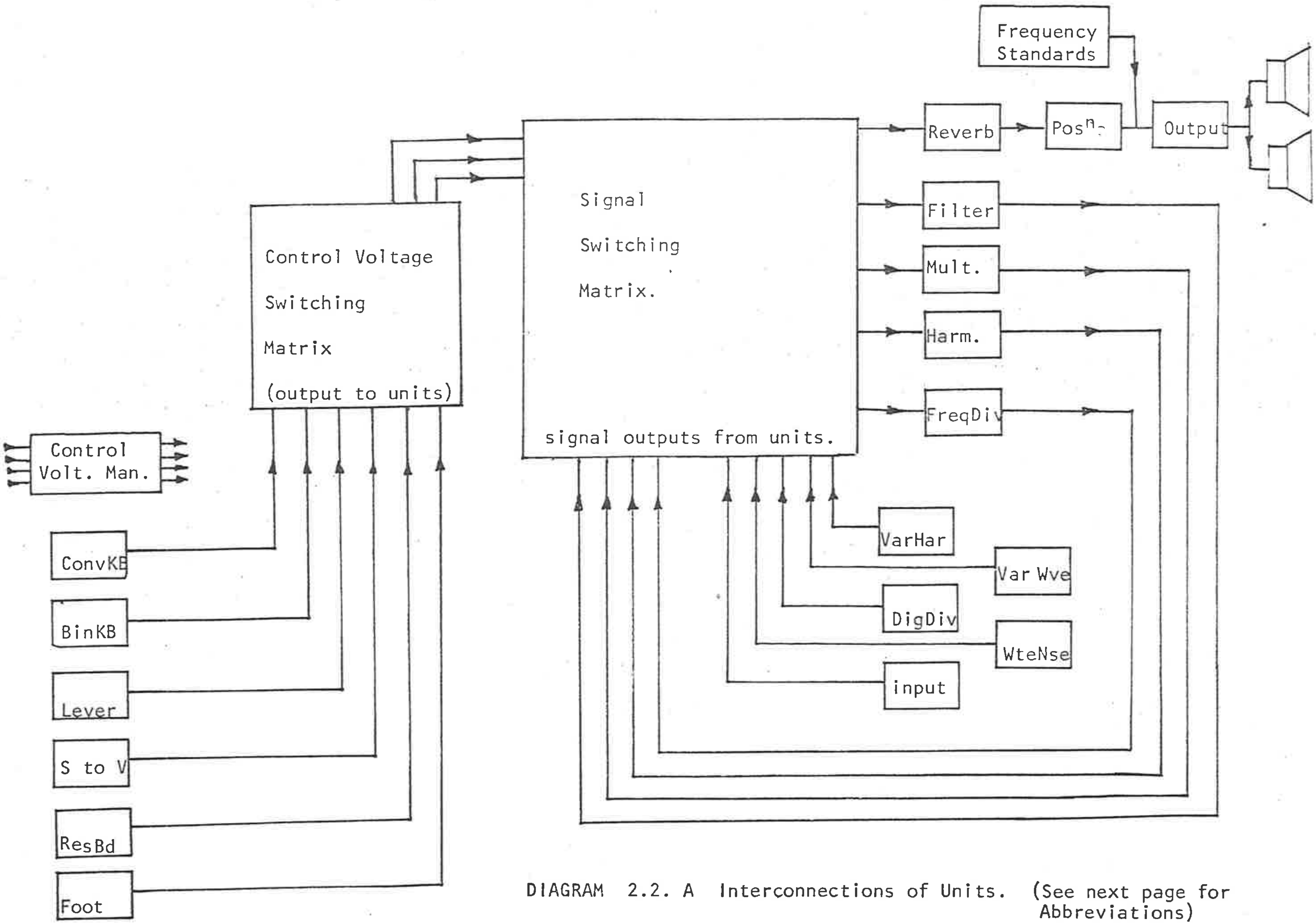


DIAGRAM 2.2. A Interconnections of Units. (See next page for Abbreviations)

Abbreviations Used in Diagram 2.2.A

Output	Output Amplifiers
Posn	Position Controller
Reverb	Reverberation Unit
Filter	Filter
Mult	Multiplier
Harm	Harmonic Generator
FreqDiv	Frequency Divider
VarHar	Variable Harmonic Oscillator
VarWve	Variable Waveform Oscillator
DigDiv	Digital Division Oscillator
WteNse	White Noise Generator
Input	Input amplifiers
ConvKB	Conventional Key Board
BinKB	Binary Keyboard
Lever	Lever Controller
StoV	Sound to Voltage Unit
ResBd	Resistance Board Controller
Foot	Foot Controller

2.3 Methods of Interconnecting the Units[1]

A significant proportion of the total resources of a synthesizer are devotec to the means of interconnecting the units. A total of 21 signal outputs are to be connected to any of 13 signal inputs, and 12 control voltages from the control units are to be connected to 13 sound sources and sound manipulators.

Patchchords enable the connection of one output to one input, but are cumbersome to use when multiple connections are to be made. Furthermore the flexing of the chords and exposure of the plugs to corrosion mean high quality and therefore expensive plugs are required to give a suitable reliability.

The cross bar switching arrangement used on telephone exahanges provides connections to many subscribers with relatively few switches. Diagram 2.3.A shows the basic cross bar switching arrangement. The number of independent lines is restricted with this arrangement. In this application the number of switches in a complete matrix of 24 x 14 for the signals and 15 x 13 for the control voltages would involve 531 switches. If all these switches were devoted to the signals alone in a cross bar switching arrangement there would be sufficient switches for only 7 independent cross bars. Cross bar switching is more suited to applications where only a small proportion of the lines are connected at any time.

The switching arrangement used in the design of the synthesizer had a signal matrix and a control voltage matrix. There is provision to obtain 3 control voltages from the signal matrix. This uses a total of 531 switches. The arrangement of switches is convenient for the musician who can trace complex switching arrangements conveniently,

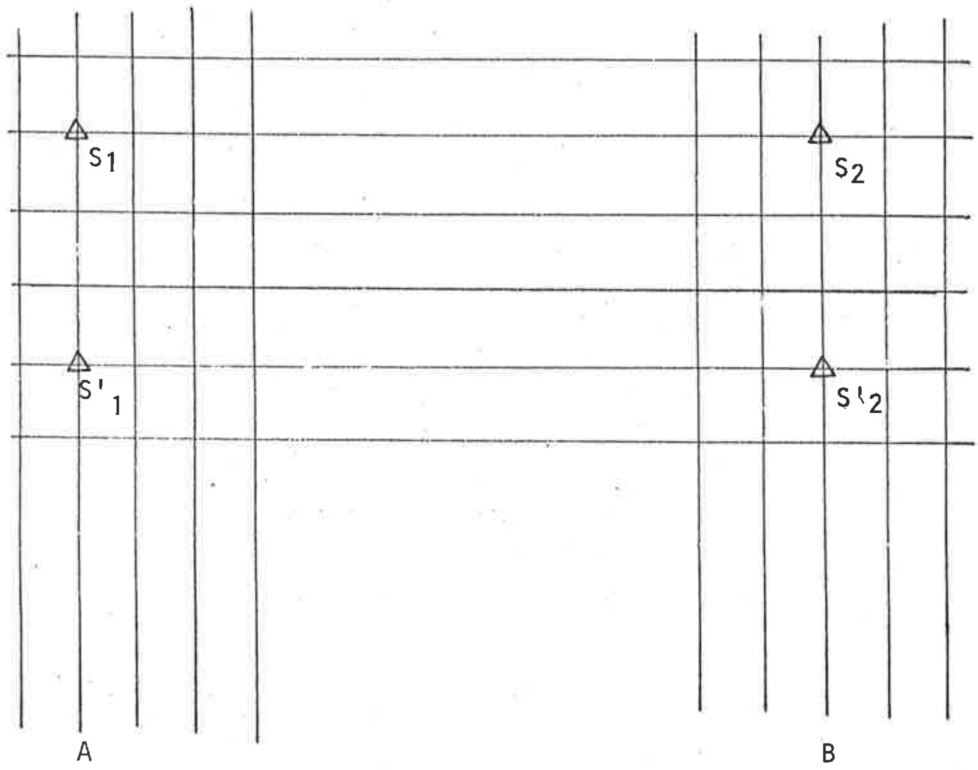


DIAGRAM 2.3.A Cross Bar Switching Arrangement. A switch is placed at the intersection of each of the lines. The connection between 'A' and 'B' is made by closing both S_1 and S_2. Another cross bar such as that involving S'_1 and S'_2 could have been used.

and can make multiple connections.

When two signal outputs are connected together the resulting signal is the average. The circuit of 2.3.B is used to provide this characteristic. This avoids overloads of the signal voltage but means when an additional signal source is switched in the amplitude of those already present is reduced.

The control switching matrix is arranged to provide the voltage sum where more than one input is used. Although this means overload's can occur and must be avoided by the operator, the setting of the voltage from one of the control units is not altered when the control voltages from other units are added. The summing circuit is given in Diagram 2.3.C.

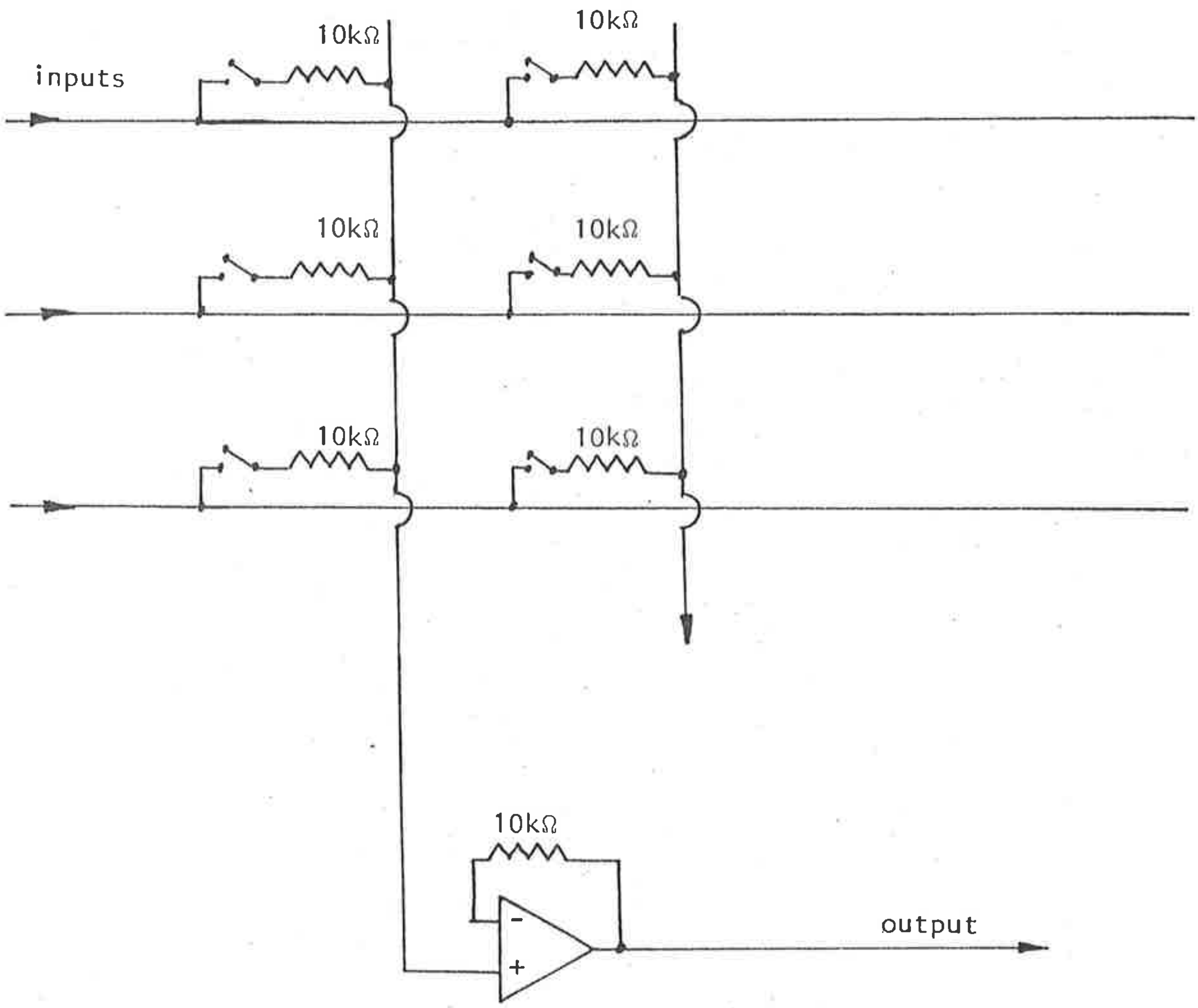


DIAGRAM 2.3.B The circuit which gives the average voltage when multiple connections are made, as with the signal switching matrix.

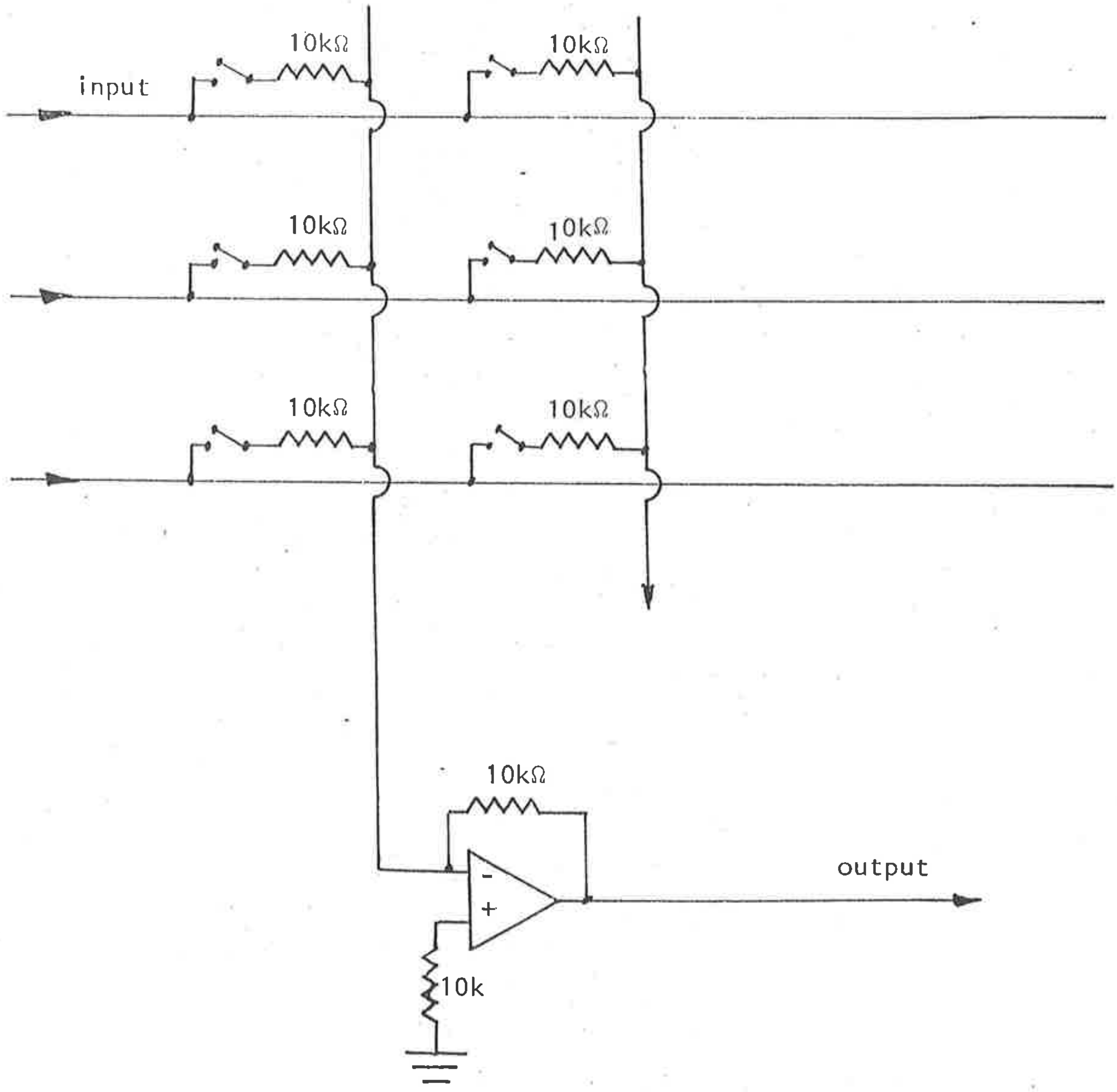


DIAGRAM 2.3.C The circuit which gives the sum of the voltages when multiple connections are made, as with the control voltage switching matrix.

NOTES FOR CHAPTER 2

1. Patchchords are used by the large Moog synthesizers and by the Buchla series 200 systems. However the ARP2500 music synthesizer and E.M.S.s' Synthi 100, VCS 3 and KB 1 synthesizers all use a matrix of pins. See reference 11 pp 79.

CHAPTER 3 SOUND SOURCES

3.1 Introduction

In this chapter the four sound source units are described. Sound sources are characterized by having only signal outputs. The design criteria are to make widely diverse types of sound outputs which can be chosen by the operator. It should be convenient for the operator to adjust the unit to give a particular sound, and to predict the type of sounds which are available.

Three of the sound sources provide periodic waveforms. The three design approaches are to combine waveforms which have distinct harmonic components, to provide a chosen waveform, and to combine submultiples of a given frequency. The remaining sound source is a white noise generator.

The Variable Harmonic Generator provides any combinations of four basic waveforms. These waveforms are chosen to have distinct harmonic contents and to be convenient to generate. These waveforms are the sinewave, the squarewave, the triangle and a double frequency sawtooth.[1]

The second oscillator approaches the selection of a periodic waveform from the time domain. It generated a waveform which is set up on 8 sliding potentiometers. These potentiometers are arranged on the front panel so that they display the eight ordinates, and therefore give a representation of the waveform. The generated waveform uses an interpolation function between the ordinates; there are 16 interpolated voltage levels between each ordinate. Zero, first, third and fifth order interpolation functions are provided.

The Digital Division Oscillator combines four square wave outputs which are obtained from the division of a high clock frequency. The divisions are by 3, 5, and 7, and also any number up to 16. The sound from this oscillator is quite unique. If the chosen divisor is 13 the fundamental frequency of the oscillator is .7 Hz, When the origional clock frequency is 1 kHz.

Each of the oscillators has two independent signal outputs. These can be used in conjunction with the signal multiplier sound manipulator to greatly enhance the range of output waveforms. When a component of the control voltage for frequency of the oscillator is another signal frequency modulation occurs. This extends the range of sounds available also.

The White Noise Generator is the only sound source which provides a non-periodic waveform. The noise component of the voltage across a reversed biased zener diode is amplified to give this white noise signal. Filtering of this noise is provided, though this is not voltage controlled. There is also a separate low frequency output which is intended as a control voltage. This gives a relatively slow random variation in the signal voltage, and can be used independently of the other signal.

The interpolated waveform of the Variable Waveform oscillator, the Digital Division Oscillator and the independent dual signal outputs are novel designs.

3.2 Variable Harmonic Oscillator

The Variable Harmonic Oscillator provides any combination of four basic waveforms, each of which have distinct harmonic contents. These waveforms are the sinewave, the triangle wave, the squarewave, and a double frequency ramp or sawtooth waveform. Each of these waveforms must be produced with a constant amplitude over a frequency range of at least 1 : 10^3. The frequency of oscillation must be voltage controlled over this frequency ratio, from 5 Hz to 5 kHz. With the design to be described a frequency ratio of 1 : 10^4 was achieved and this enables a very low frequency of oscillation. This means the oscillator can be used to provide a low frequency control voltage. This frequency ratio compares favourably with the recently available integrated circuit voltage controlled oscillators which have a frequency range of 1 : 10^3 with one sweep.

The sinewave is important musically because it is the only waveform which has no harmonic components. The sinewave can be obtained by frequency independent waveshaping of a triangular waveform. The magnitude of this triangular waveform must be constant at all the operating frequencies. A simple relaxation oscillator which gives both a triangular wave and a square wave which both have constant amplitude is shown in diagram 3.2.A. During each cycle the integrator reaches the trigger level of the schmitt trigger which then resets the input voltage to the integrator. The integrator output then drops to the trigger level of the schmitt trigger and the cycle is repeated. The frequency of this oscillator can be controlled using an analog inverter as shown in diagram 3.2.B. In this form the oscillator has a linear control voltage to frequency relationship.

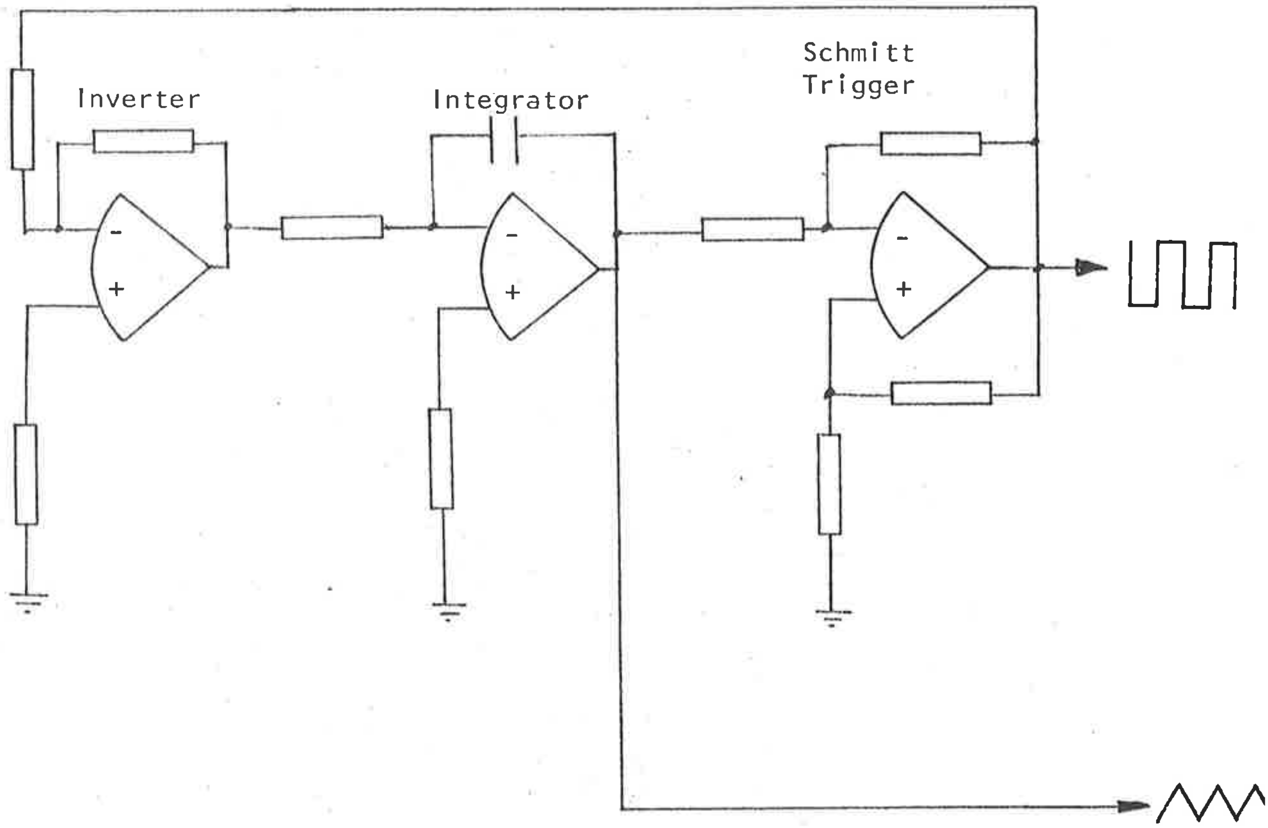


DIAGRAM 3.2.A A relaxation oscillator which gives triangular and square waveforms.

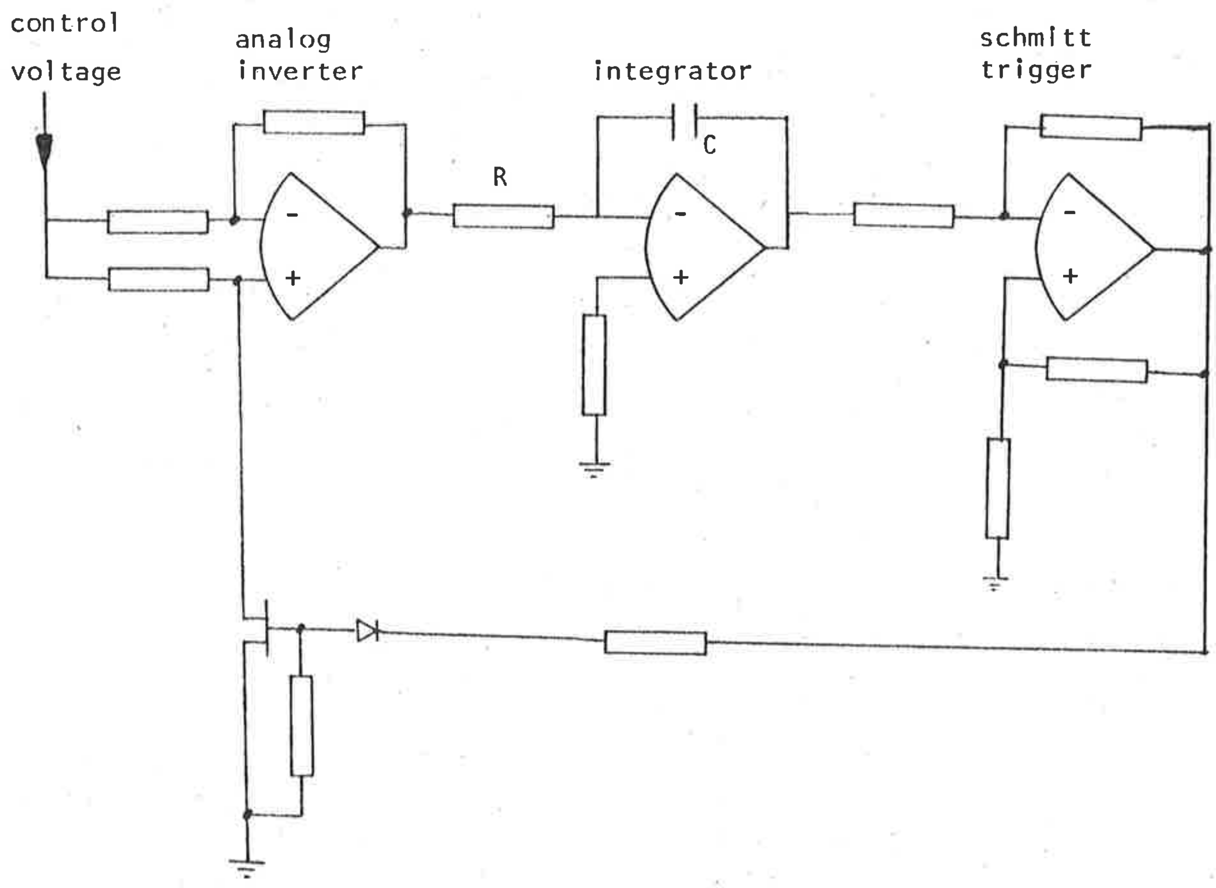


DIAGRAM 3.2.B Voltage control of the frequency of oscillation of the relaxation oscillator of diagram 3.2.A using an analog inverter.

The frequency of the oscillator is determined by the rate of charge of the voltage across the capacitor, C of diagram 3.2.B. In turn this is determined by the current through the resistor R, if the current to the inverting input of the operational amplifier is neglected. If this resistor, R, is linear the voltage across it is proportional to the current through it, and hence this voltage is proportional to the frequency of the oscillator. The maximum voltage across the resistor is ±15 V because of the voltage limitations of the output voltage on the integrated circuit, The voltage across this linear resistor which would give a frequency 10^3 times lower than the maximum frequency is 15 mV. At this voltage level the voltage offsets which are present at the input of the integrated circuit are significant. The precision linear operational amplifier, the μA 777 has a maximum voltage offset of ±3 mV. The effect of these input voltage offsets of the integrator and analog inverter is to lengthen the time taken for one half cycle, and to reduce the time for the other half cycle. This means all the waveforms of the oscillator are distorted at low frequencies. Offset currents to the integrator have a similar effect on the waveforms but in this case the extent of this effect on the waveforms can be minimized by a suitable choice of the resistor R and the capacitor C which form the integrator. It is possible to ensure that the charging currents are large in relation to the offset currents. However the input offset voltage provide a limiting factor because of the maximum voltage swing across the resistor R.

The frequency ratio of the oscillator can be considerably extended using a nonlinear resistive element. The desired control voltage to frequency ratio is logarithmic; a given increment in the control voltage should increase the frequency by a given ratio. If the corresponding non-linear element in introduced at R the ratio of the currents through this resistor can greatly exceed the ratio of the corresponding voltage across it. This removes the previous restriction the input voltage offsets of the integrated circuits places on the maximum frequency ratio.

The symmetry of the generated waveforms depend upon the symmetry of this nonlinear resistive element. It is important that the voltage to current relationship across it is an odd function; $i(v) = -i(-v)$. If diode shaping for the positive applied voltages is realized independently of the diode shaping for the negative applied voltages, it would be necessary to match these two diode shaping networks quite accurately so that the symmetry of the generated waveform was not impaired. The use of a bridge rectifier circuit of diagram 3.2.C ensures a very accurate matching of the forward and the reverse characteristics of the nonlinear element. The bridge consists of matched diodes, and the voltage applied to the remainder of the nonlinear element is always in the same direction.

The resistors in the nonlinear resistor are chosen to give a logarithmic relationship between voltage and current. These resistors shunt diodes which are chosen to have equal forward voltage drop. It can be shown that the resistance of the resistors should be in a geometric progression if the desired logarithmic control voltage to frequency characteristic is to be obtained.

The desired relationship is $i = k_1 e^{v + k_2}$ 3.2.i

where i is the current through the nonlinear element

v is the voltage across it

and k_1 and k_2 are constants.

It follows that $di/dv = k_3 i$ where k_3 is also a constant. 3.2.ii

When successive resistors in the nonlinear element are in a geometric progression with the ratio of r between adjacent resistors (the resistors can be labelled so that $r < 1$) then the incremental resistance of arbitrarily small currents is:

$$dv/di = R + rR + r^2R + r^3R + r^4R + \ldots r^mR \ldots\ldots$$

When the current i is such that the voltage drop across a diode exceeds .6 V then the incremental resistance of this section becomes zero. In this case the incremental resistance of the total nonlinear network has become:

$$dv/di = r^n(R + rR + r^2R + r^3R + r^4R \ldots r^{m-n}R \ldots)$$

where the n^{th} diode is fully conducting.
If the current level were to increase by a factor of $k_3 = r^{-i}$ then the incremental resistance would become;

$$dv/di = r^{n+i}(R + rR + r^2R + r^3R + \ldots r^{m+i-n}R \ldots)$$

This means that if the current is increased by r^{-i} the incremental conductance is also increased by $k_3 r^{-i}$, and the relationship 3.2.ii is satisfied.

There will ofcourse be a finite number of series elements. The value of the final resistor will be given by:

$$R_{max} = r^{n_{max}} (R + rR + r^2R + r^3R \ldots\ldots)$$
$$= r^{n_{max}}R/(1-r)$$

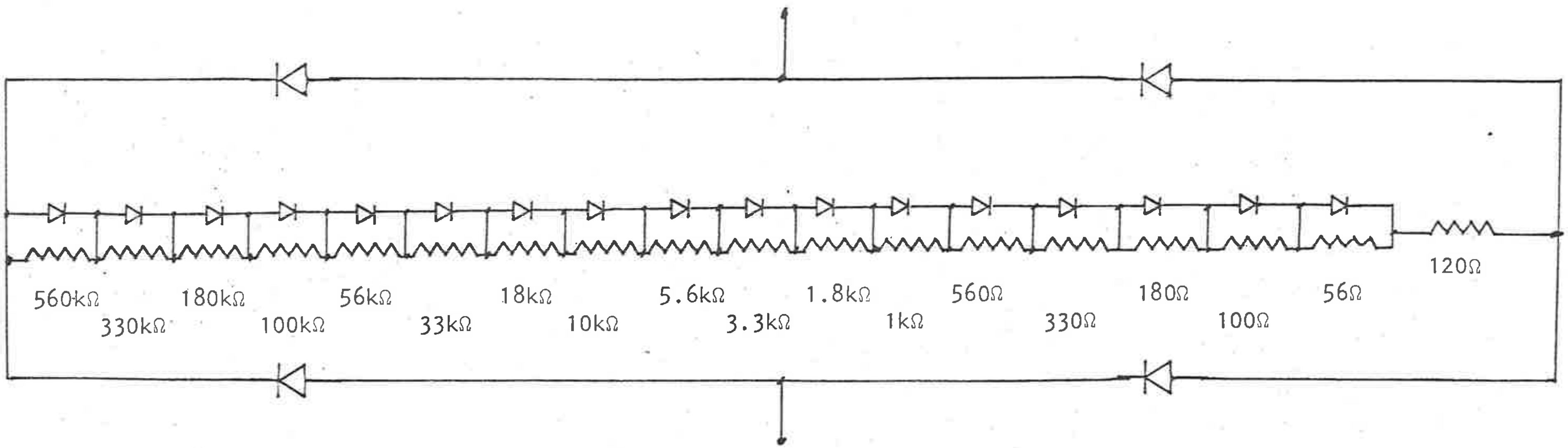


DIAGRAM 3.2.C. The nonlinear resistor R which is composed of a bridge circuit and a series of resistors which are in a logarithmic progression.

A linear resistor of this value in series with the nonlinear series elements will ensure that the logarithmic relationship will be satisfied over the range of interest. The value of this resistor is chosen to correspond to the highest operating frequency.

The significance of this design approach of a logarithmic voltage to current resistive element is that readily available resistors can be used because the preferred values are in a geometric progression.

The harmonic components of the waveforms produced by the oscillator are given in diagram 3.2.D. The harmonic components of each of these decreases with frequency. The square wave has all the odd harmonics, with the n^{th} harmonic having a magnitude of 1/n times the magnitude of the fundamental sinusoidal component. The double frequency ramp has all the even harmonic frequencies of the fundamental oscillator frequency. Again the magnitudes of these components decreases as the order of the harmonic. The triangle waveform has all the odd harmonics, but in this case the magnitude of the n^{th} harmonic is $1/n^2$ times the magnitude of the fundamental component.

The harmonic components of all these waveforms decrease with frequency. The Harmonic Generator sound manipulator can be used in conjunction with the triangular of ramp waveforms to give predominantly higher frequency harmonics.

The block diagram of this oscillator is given in diagram 3.2. E, and the circuit diagram is given in diagram 3.2.F.

Two independent signal output are available to the musician, and two voltage controlled attenuators are provided. These attenuators use complementary field effect transistors so that one signal output increases

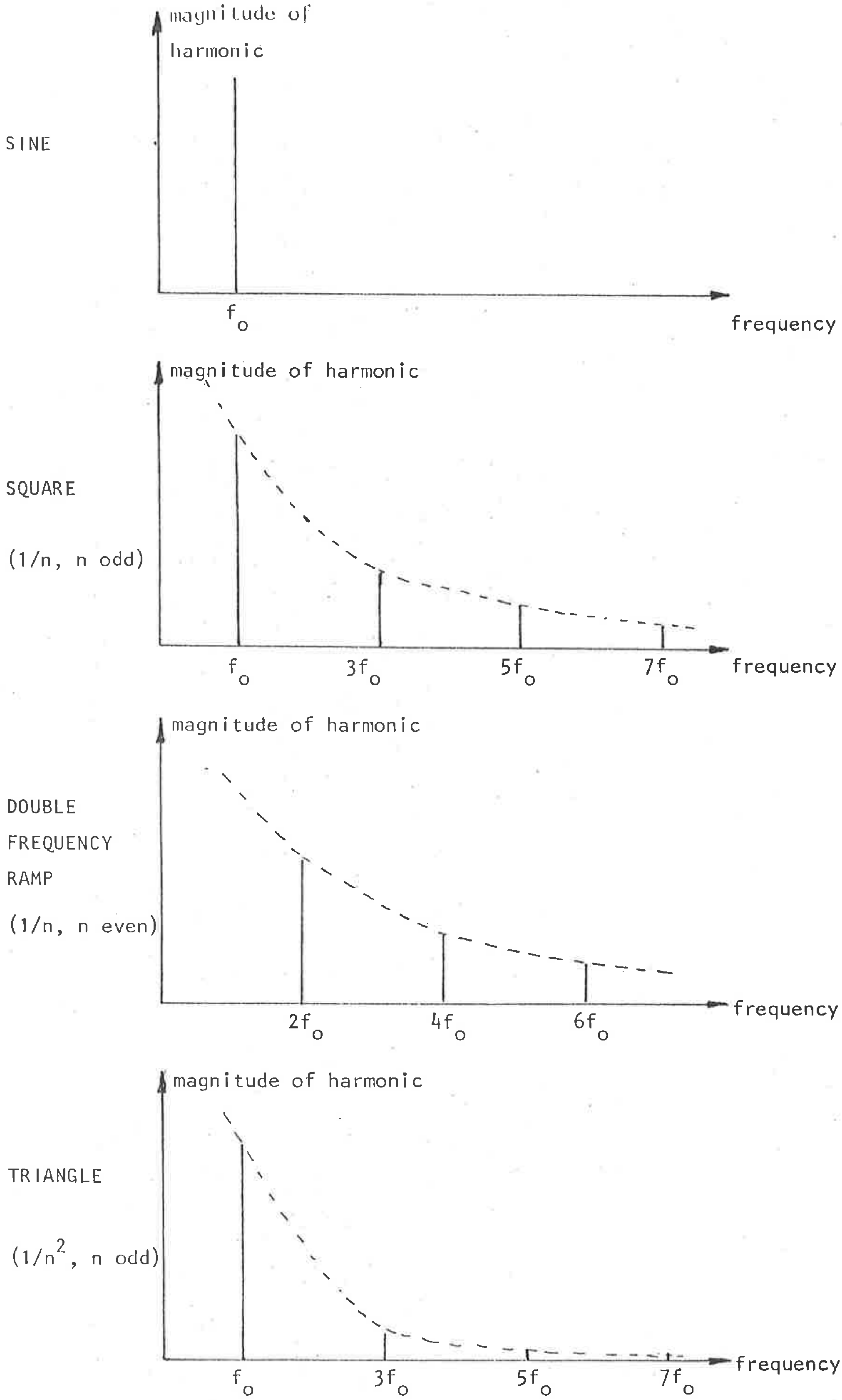


DIAGRAM 3.2.D. Harmonic components produced by the oscillator.

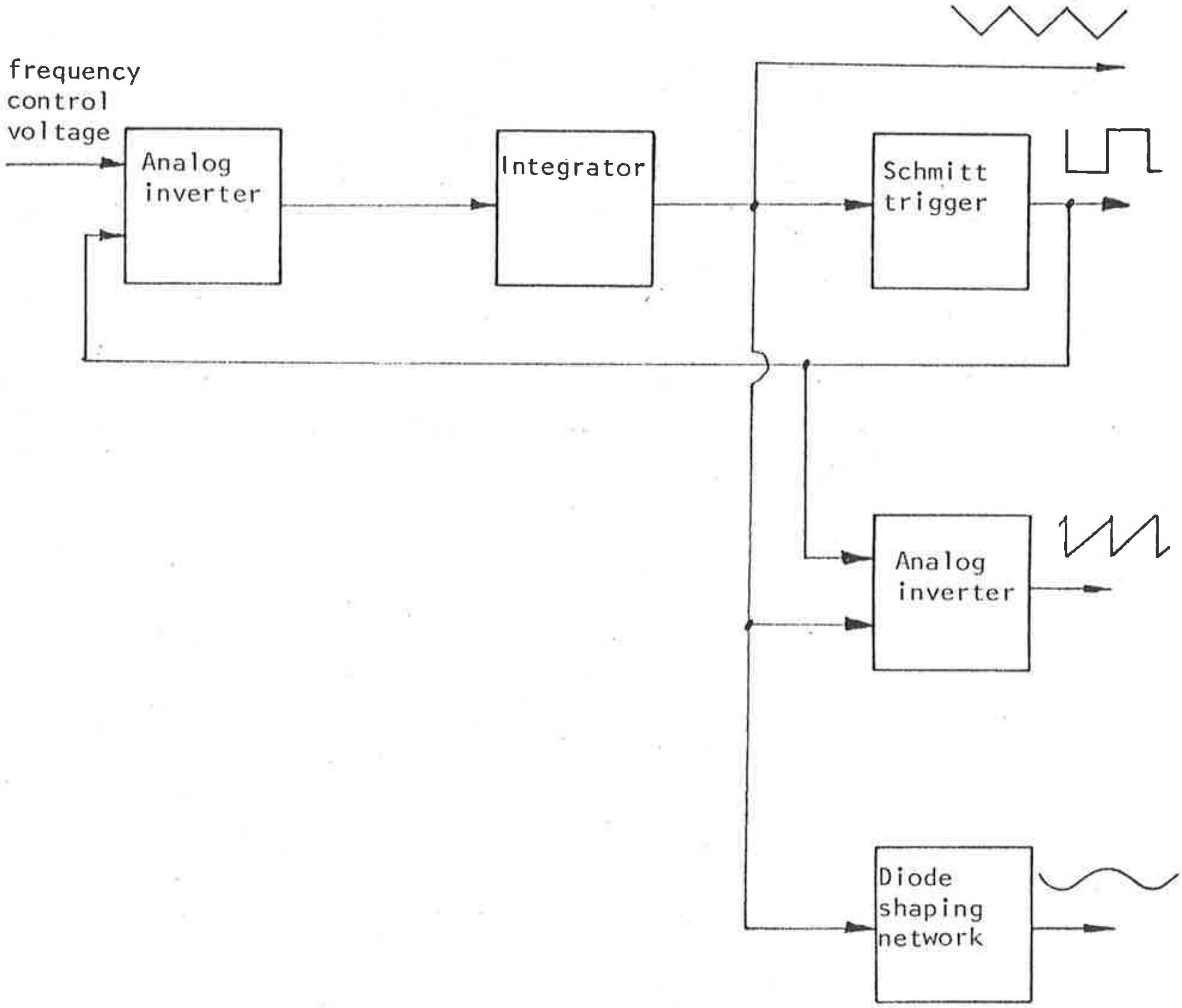


DIAGRAM 3.2.E. Block diagram of the variable harmonic oscillator

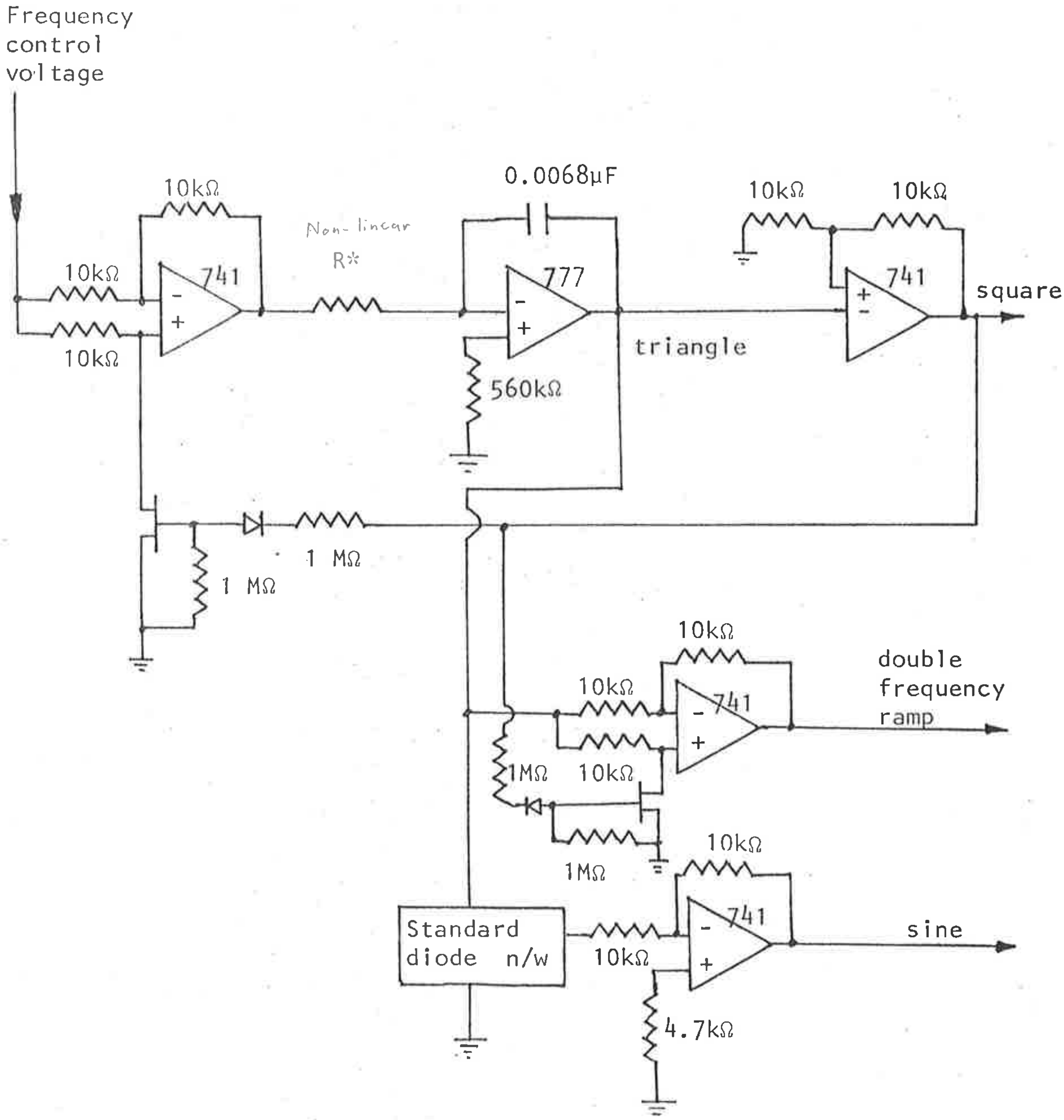


DIAGRAM 3.2.F. Circuit diagram of the variable harmonic oscillator.

* See diagram 3.2.C.

as the other signal output decreases when the same control voltage is used. This is achieved without using an operational amplifier to invert the amplitude control voltage. Two trim potentiometers for each field effect transistor allow a centre point and slope adjustment of the control voltage so that the characteristics of the field effect transistors can vary widely. The signal level across the field effect transistor is minimal so there is minimum distortion of the signal. Diagram 3.2.G gives this circuit.

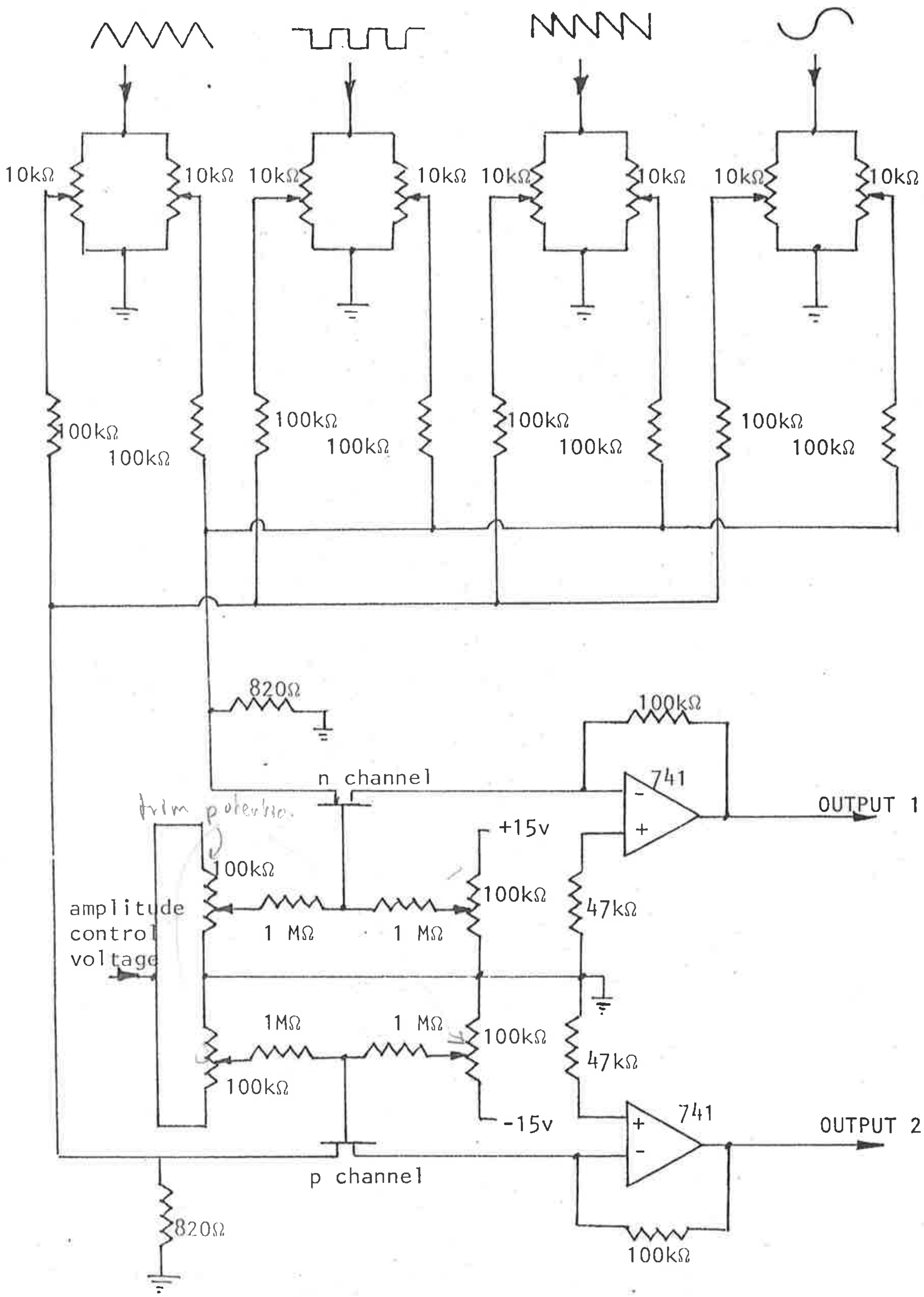


DIAGRAM 3.2.G. Attenuator section of the variable harmonic oscillator.

3.3 Variable Waveform Oscillator

The variable waveform oscillator enables the operator to choose the sound of a periodic signal by directly adjusting the waveform of the signal. This is the time domain approach to the selection of a periodic signal, as distinct from the frequency domain approach of the variable harmonic oscillator. The purpose of providing different methods of selecting the sound is to provide differing types of sound for the musician.

The problem of providing an adjustable waveform which is independent of frequency could now be solved using one digital memory element. However, when digital storage was first investigated for this oscillator the cost was prohibitive. An alternative approach used in analog computers for non-linear computations is to use a diode function generator. The gradients of segments of waveforms are adjusted, and provision must be made for both positive and negative slopes: Simpler diode function generators only provide monotonic input to output voltage relationships.

An alternative approach used in earlier analog computers was to use a sheet of cardboard which had the desired waveform cut along the horizontal edge. This is inserted between an oscilloscope and a light detector. A control loop is used to determine the vertical height of the spot so as to keep the spot half covered by the cardboard. The timebase can then be set at an independent frequency, and the generated waveform is the same at all these frequencies. One disadvantage of this method is that pulses would occur on the waveform during the reset scan. It is important to avoid any impulses on the waveform in this application because these higher frequency components are particularly conspicuous to the musician. The established technique could be changed slightly so that a circular rotation of the spot was used. In this case there would be no discontinuity at the end of each cycle. The advantage of providing a visual image of the waveform is that the operator can use this to help set up a sound. However in this case it would be necessary for the operator to make another cardboard cross-section each time the waveform is to be changed.

The oscillator built for the synthesizer provides a display of the waveform and allows continuous adjustment which is monitored. Eight sliding potentiometers were used to adjust eight ordinates of the waveform, and these are arranged on the front panel of the oscillator so that there is a display of the waveform.

The oscillator's output is switched between these eight ordinates in sequence. However if the simple system of switching from one ordinate to the next were used there would always be a dominant eighth harmonic present in the sound of the oscillator. It is necessary to interpolate between these ordinates to avoid such a characteristic sound.

With this interpolation it is necessary to provide a circuit which changes its output uniformly from one voltage level to another in a given time. The use of integration was considered, but it was found that there could be difficulties in maintaining the required waveform over a large range of frequencies. The use of a non-linear resistive element to achieve this frequency range, as with the Variable Harmonic Oscillator (3.2.) was not possible because of the many adjustable slopes which made up the waveform.

A time independent digital interpolation was used. Digital switching is used to obtain seven intermediate voltage levels between each pair of the eight ordinates. This means there are 64 discrete voltage levels during each cycle, and zero, first and fourth order interpolations are provided. This means the characteristic sound of the oscillator can be mellow or harsh.

The technique used in this oscillator was to use analog switches to combine the outputs from the eight ordinates in proportions which are determined by the interpolation function. With the linear interpolation function only two ordinates are combined at any intermediate point. The voltage level which is midway between the ordinates is the average of these two ordinates. Similarly, voltage levels whcih occur closer in time to one of the ordinates are given by the weighted average.

Diagram 3.3.A. gives the desired output from the oscillator in terms of the linear interpolation between two ordinates p_1 and p_2. The equation for the output voltage for the linear interpolation is;

$$v(t) = p_1 + \frac{p_2 - p_1}{\Delta t} . t$$

where Δt is the time interval between the desired occurrence of successive ordinates p_1 and p_2. The equation applies only within this time interval.

The circuit realization of this lineat interpolation function is given in diagram 3.3.B. Three intermediate voltage levels are provided at time intervals of $\Delta t/4$. For the n^{th} of these time intervals;

$$v(n) = p_1 + (p_2 - p_1)n/4$$

$$= p_1(1 - n/4) + p_2 n/4$$

This equation express the relative weighting of the ordinates in terms of the number of the intermediate ordinate, n. Table 3.3.A. gives this relative weighting, and table 3.3.B. gives the corresponding resistors in the circuit of Diagram 3.3.B. which provide this interpolation function.

One further linear interpolation was made by the oscillator to provide a total of seven interpolated voltage levels. Diagram 3.3.D. gives the complete switching diagram of the oscillator.

This technique used for the linear interpolation can now be extended to higher order functions. It is necessary to consider two ordinates for a first order interpolation, and similarly it is necessary to consider five ordinates for the fourth order interpolation.

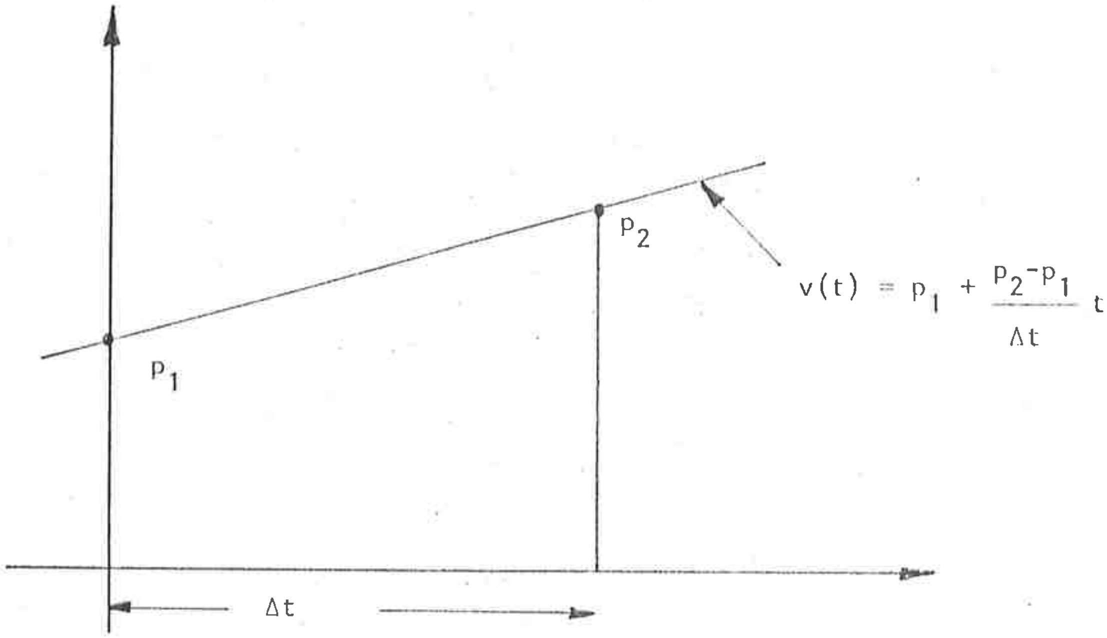


DIAGRAM 3.3.A. Algebraic expression for the ideal output voltage from the oscillator for a linear interpolation

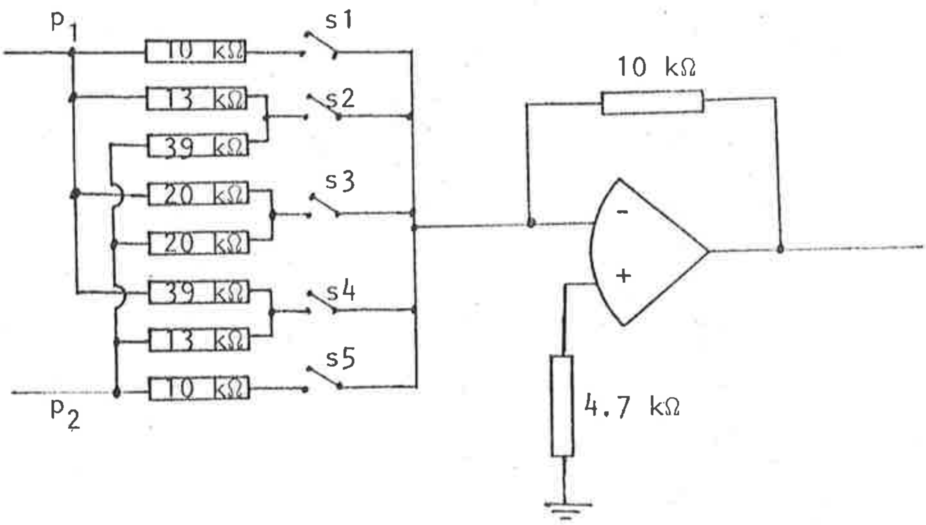


DIAGRAM 3.3.B. Use of analog switches to obtain three interpolated voltage levels.

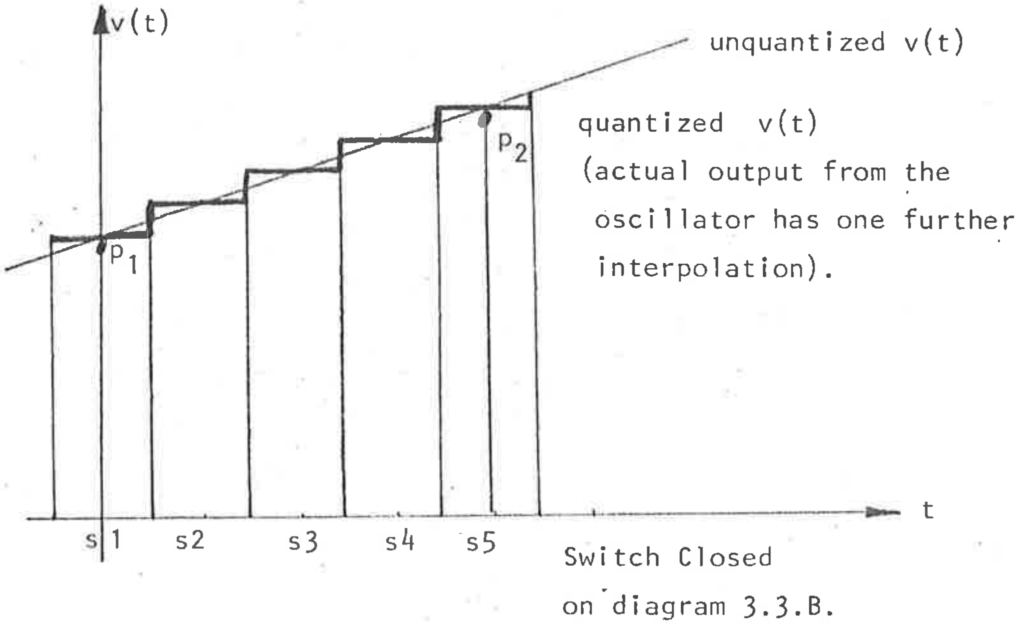


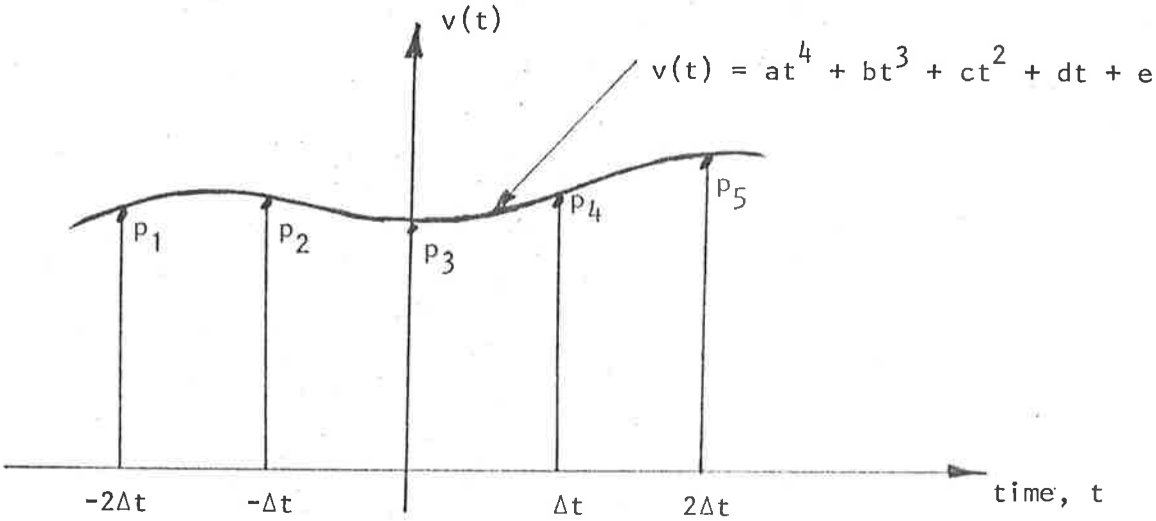
DIAGRAM 3.3.C. The output voltage of the oscillator for four linear interpolations.

Number of Interpolation n	Ordinate p_1	Ordinate p_2
0	1	0
1	.75	.25
2	.5	.5
3	.25	.75
4	0	1

TABLE 3.3.A. Relative weighting of ordinates p_1 and p_2 for a linear interpolation.

Number of Interpolation n	Ordinate p_1	Ordinate p_2
0	10 kΩ	∞
1	13.3 kΩ	37.5 kΩ
2	20 kΩ	20 kΩ
3	37.5 kΩ	13.3 kΩ
4	∞	10 kΩ

TABLE 3.3.B. Resistors required in the circuit of Diagram 3.3.B. to provide the interpolation function.



$$P_1 = 16\,a\,t^4 - 8\,b\,t^3 + 4\,c\,t^2 - 2\,d\,t + e$$

$$P_2 = a\,t^4 - b\,t^3 + c\,t^2 - d\,t + e$$

$$P_3 = e$$

$$P_4 = a\,t^4 + b\,t^3 + c\,t^2 + d\,t + e$$

$$P_5 = 16\,a\,t^4 + 8\,b\,t^3 + 4\,c\,t^2 + 2\,d\,t + e$$

DIAGRAM 3.3.D. A fourth order interpolation function between five ordinates. The equations relate the value of the interpolation function to the ordinates at the required time.

Diagram 3.3.D. gives the five ordinates p_{1-5} in terms of the order of the polynomial to be fitted to them. These five equation expressing these ordinates in terms of the polynomial can be used to solve for the coefficients of t in terms of the ordinates. It can be shown that;

$$a = (p_1 - 4p_2 + 6p_3 - 4p_4 + p_5)/(24\Delta t^4)$$

$$b = (-p_1 + 2p_2 - 2p_4 + p_5)/(12\Delta t^3)$$

$$c = (-p_1 + 16p_2 - 30p_3 + 16p_4 - p5)/(24\Delta t^2)$$

$$d = (p_1 - 8p_2 + 8p_4 - p5)/(12\Delta t)$$

and $e = p_3$

where the five ordinates are to be fitted by the fourth order polynomial in t;

$$v(t) = at^4 + bt^3 + ct^2 + dt + e \qquad \ldots\ldots \qquad 3.3.i$$

As was the case with the linear interpolation the voltage levels at three intermediate time intervals can be calculated in terms of the five ordinates p_{1-5}. Eliminating the coefficients a, b, c, d, and e, from the defining equation 3.3.i. gives;

$$\begin{aligned} v(n) = \; & p_1(n^4/24 - n^3/12 - n^2/24 + n/12) \\ + \; & p_2(-n^4/6 + n^3/6 + 2n^2/3 - 2n/3) \\ + \; & p_3(n^4/4 - 5n^2/4 + 1) \\ + \; & p_4(-n^4/6 - n^3/6 + 2n^2/3 + 2n/3) \\ + \; & p_5(n^4/24 + n^3/12 - n^2/24 - n/12) \end{aligned}$$

For the three interpolated voltages n = 1/4, 2/4, and 3/4 respectively.

Substituting n into these equations gives the weighting factors of the five ordinates in order to obtain the fourth order interpolation function. Table 3.3.C. gives these factors. The corresponding values of the resistances which are required to give these weighting factors is given in table 3.3.D.

Number of Interpolation	n	Ordinates p_1	p_2	p_3	p_4	p_5
0	0	0	0	1	0	0
1	1/4	.017	-.123	.923	.205	-.021
2	2/4	.023	-.156	.703	.649	-.039
3	3/4	.171	-.107	.375	.752	-.038
4	1	0	0	0	1	0

TABLE 3.3.C. Relative weighting of ordinates p_{1-5} for a fourth order interpolation.

Number of Interpolation	n	Ordinates p_1	p_2	p_3	p_4	p_5
0	0	∞	∞	10kΩ	∞	∞
1	1/4	590kΩ	81 kΩ	11 kΩ	49 kΩ	460kΩ
2	2/4	430kΩ	64 kΩ	14 kΩ	21 kΩ	250kΩ
3	3/4	580kΩ	93 kΩ	27 kΩ	13 kΩ	270kΩ
4	1	∞	∞	∞	10 kΩ	∞

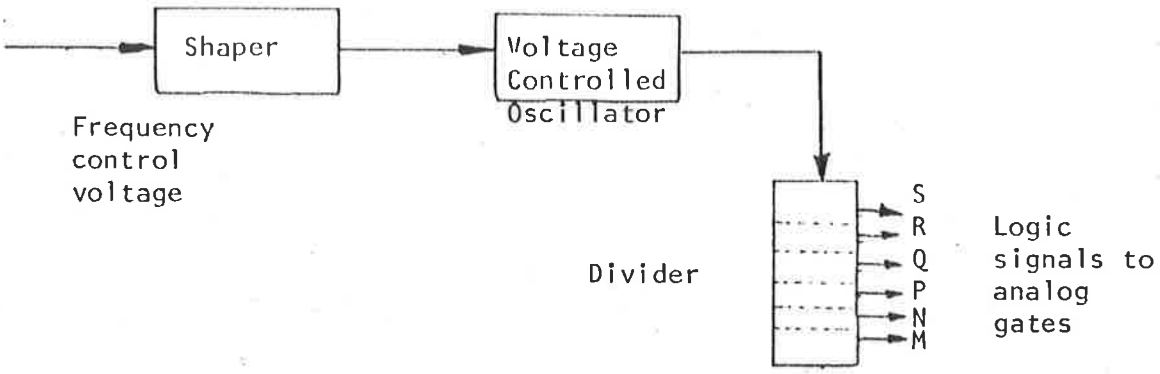
TABLE 3.3.D. Resistors required to provide the fifth order interpolation function.

The resistance matrix formed from tables 3.3.B for the linear interpolation and table 3.3.D for the fourth order interpolation were connected to an edge connector which was plugged into the oscillator as required. This meant the order of the interpolation function was changed by replacing this resistance matrix.

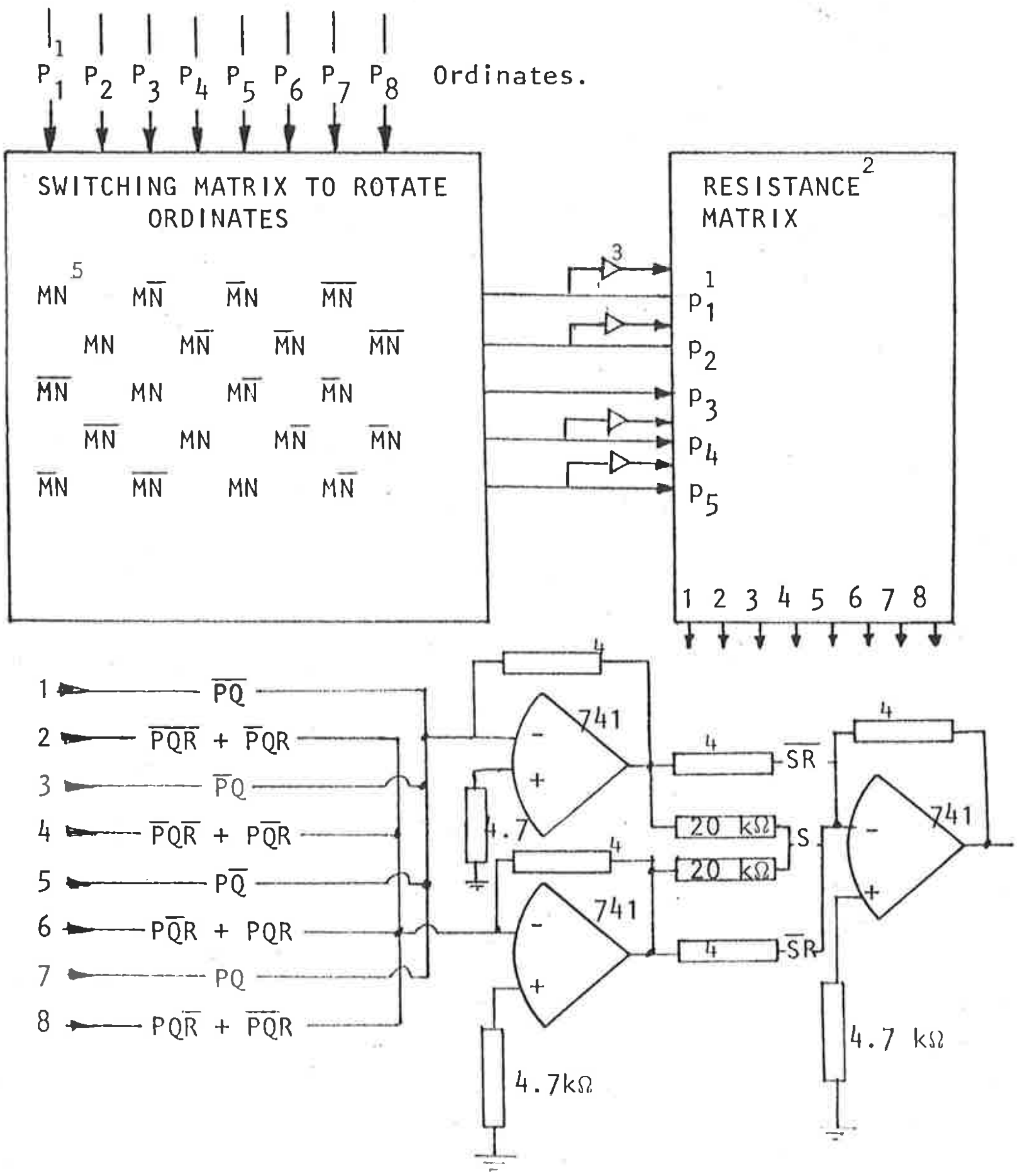
The resistors were calculated for three intermediate values only. Another linear interpolation between these voltage levels gives a total of seven intermediate values between each pair of ordinates and 64 ordinates for each cycle. Over the small interval between two successive ordinates the linear interpolation is sufficiently accurate. This technique reduces the number of resistors and analog gates required to almost one half.

In order to further reduce the number of analog switches, each fourth order interpolation function was made over two of the eight ordinates, and only then was the ordinates to the resistance matrix switched. This reduced the number of switches required because the ordinates fed into the resistance matrix need be changes only 4 times instead of 8. However this saving of switches is somewhat offset by the additional number of switches required to do the interpolation between two ordinates instead of only one.

As with the Variable Harmonic Oscillator there are two separate outputs from this oscillator. The second output is taken from a tone control system which is adjusted on the front panel of the oscillator. Diagram 3.3.E. gives the circuit description of the Variable Waveform Oscillator.



Logic Section



Switching Section

DIAGRAM 3.3.E. Switching Functions for the Variable Harmonic Oscillator.

(See next page for notes).

Explanation of Diagram 3.3.E

The logic section of the variable harmonic oscillator includes a voltage controlled oscillator.

A control voltage shaping network is used to give a logarithmic relationship between the frequency of the oscillator and the corresponding control voltage. The output frequency of the voltage controlled oscillator is divided by logic circuitry in six sections; each section giving a division by 2. The six logic signals from this divider unit, S R Q P N and M, operate analog switches. These switches determine the output voltage of the variable harmonic oscillator, by providing interpolations between 8 voltage levels. These 8 voltage levels are set up using potentiometers as ordinates on the front panel of the oscillator. The analog switches operate in three sections.

Section 1

The most significant bits from the division, M and N, are used to switch the voltage from the 8 potentiometers in the front panel. Five of these voltages are selected at any given time and fed into the resistance matrix. This matrix provides 8 interpolations between these voltage levels. A resistance matrix providing a fifth order interpolation was used. Once the least significant bits P Q R and S have switched between the interpolations, the 5 input voltages to the resistance matrix are rotated.

Section 2

The switches controlled by the logic signals P Q and R are used to switch between the 8 interpolations which are provided by the resistance matrix. Two ouputs are provided by this section. Interpolations 1, 3, 5 and 7 are present in succession on the upper 741 amplifier, whilst interpolations 2, 4, 6 and 8 are present on the lower 741 amplifier.

Section 3.

The least significant bits, D and R, are used to provide a linear interpolation between the two outputs of section 2. Of the least significant bit, S = 1, the output of the oscillator is the average of the two ouputs provided by section 2. If S = 0 and R = 0 the output of the oscillator is given by the output of the upper 741 amplifier (and the lower 741 amplifier is being switched). Conversely, if S = 0 and R = 1 the lower 741 amplifier determines the output of the oscillator, and the upper 741 amplifier is being switched.

Notes for Diagram 3.3.E.

1 Upper case 'P' denotes the ordinate which is set up by the potentiometer on the front panel. Lower case 'p' denotes the ordinate which is fed into the resistance matrix

2 The resistance matrix is given by table 3.3.B. for the linear interpolation, and by table 3.3.D. for the fourth order interpolation.

3 These inverters provide the sign reversal which is necessary for the fourth order interpolation.

4 All these resistors are 10 kΩ.

5 This switch is conducting when M = logic 1 and N = logic 1. Similarly the switch $M\overline{N}$ is conducting when M = logic 1 and $\overline{N}$ is logic 1 (N is at logic 0).

When listening to the sound of the oscillator improvements to the design become apparent. With the present oscillator, if every second sliding potentiometer is altered the fourth harmonic is readily obtained. Similarly if the sliding potentiometers are moved in groups of two then the second harmonic is obtained. It is relatively difficult to set up the oscillator to that the third harmonic is obtained though, with practice, this can be done. If 12 ordinates had been provided by the oscillator the second, third, fourth and sixth harmonics would be readily available by moving the appropriate combinations of sliding potentiometers.

The provision of only the first order interpolation would greatly simplify the oscillator, and it was found that the reduction of harmonics using the fifth order interpolation was only marginal. The acoustic effect was less important than was originally expected, although the voltage waveforms were much smoother on the oscilloscope.

3.4 Digital Division Oscillator

The Digital Division Oscillator obtains its signal output by dividing a high clock frequency. This approach is suggested by the convenience of frequency division using digital circuitry. In turn this approach is related to the design criteria of the synthesizer. A particular class of sounds can be readily generated using electronic equipment. In this way the art form reflects the electronic medium.

The design criteria suggest a wide variety of sound outputs should be provided. To ensure this the frequency divisions are by the smallest prime numbers, 2, 3, and 5. There is also division by an adjustable number $N < 16$. The four outputs from these division processes are added to give the signal output. The amplitude of each of these components is independently adjustable.

Whereas the previous oscillators have provided frequency components which are multiples of a controlled frequency, this multiplier provides submultiples also. The difference can best be illustrated when the adjustable number $N = 13$. In this case a clock frequency of 1 kHz gives a fundamental frequency of 2.6 Hz. This is well below the audio spectrum and gives this oscillator a distinctive sound.

Diagram 3.4.A gives the block diagram of the oscillator, and diagrams 3.4.B to 3.4.E give the circuit.

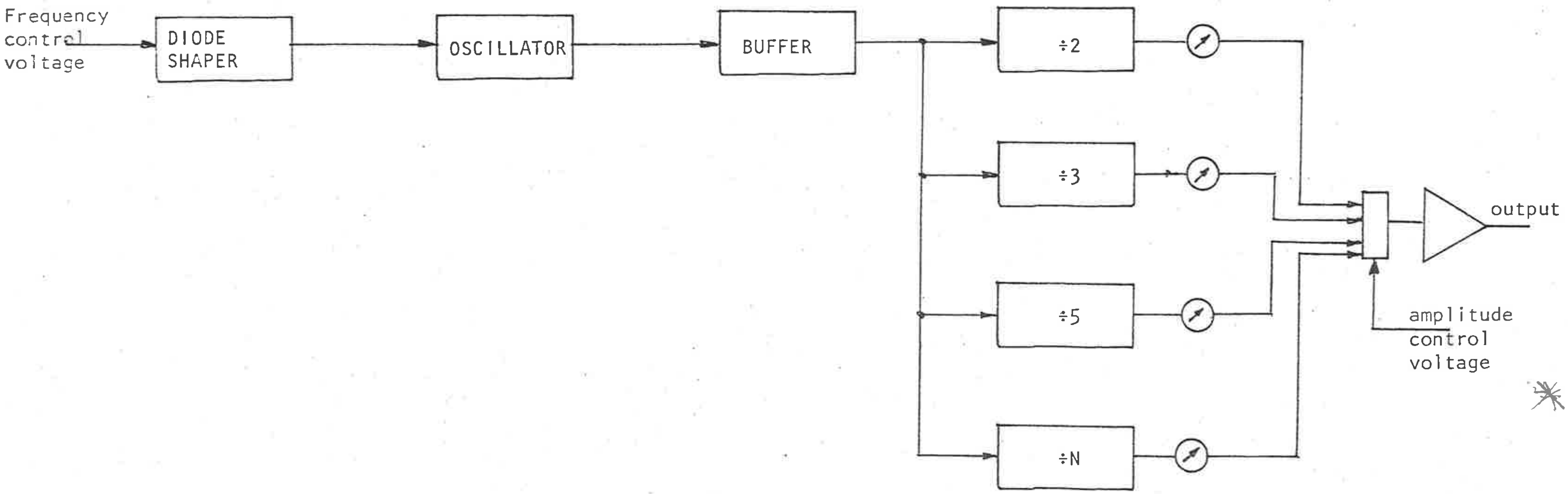


DIAGRAM 3.4.A. Block Diagram of the Digital Division Oscillator.

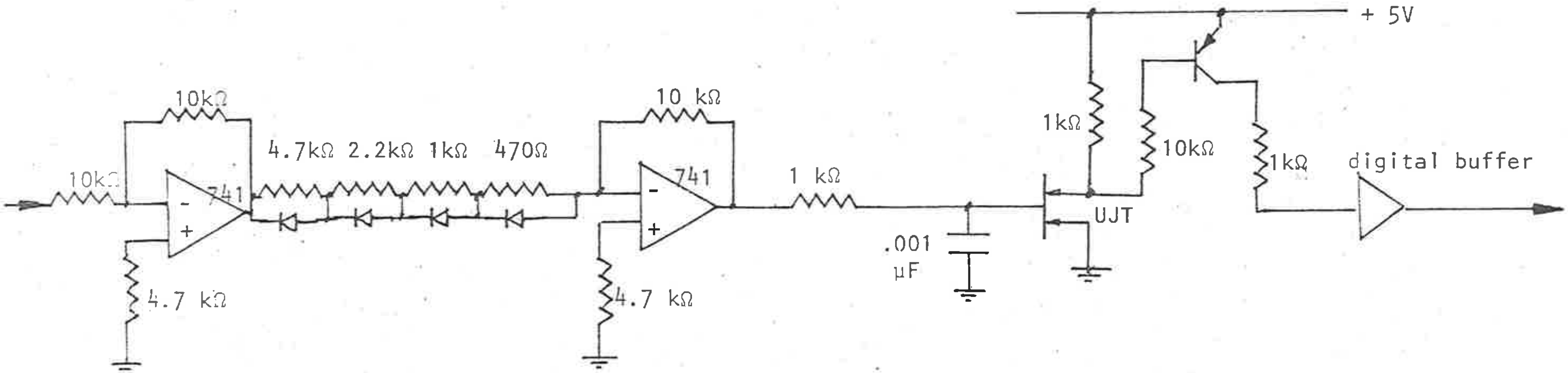


DIAGRAM 3.4.B. Oscillator section of the digital division oscillator.

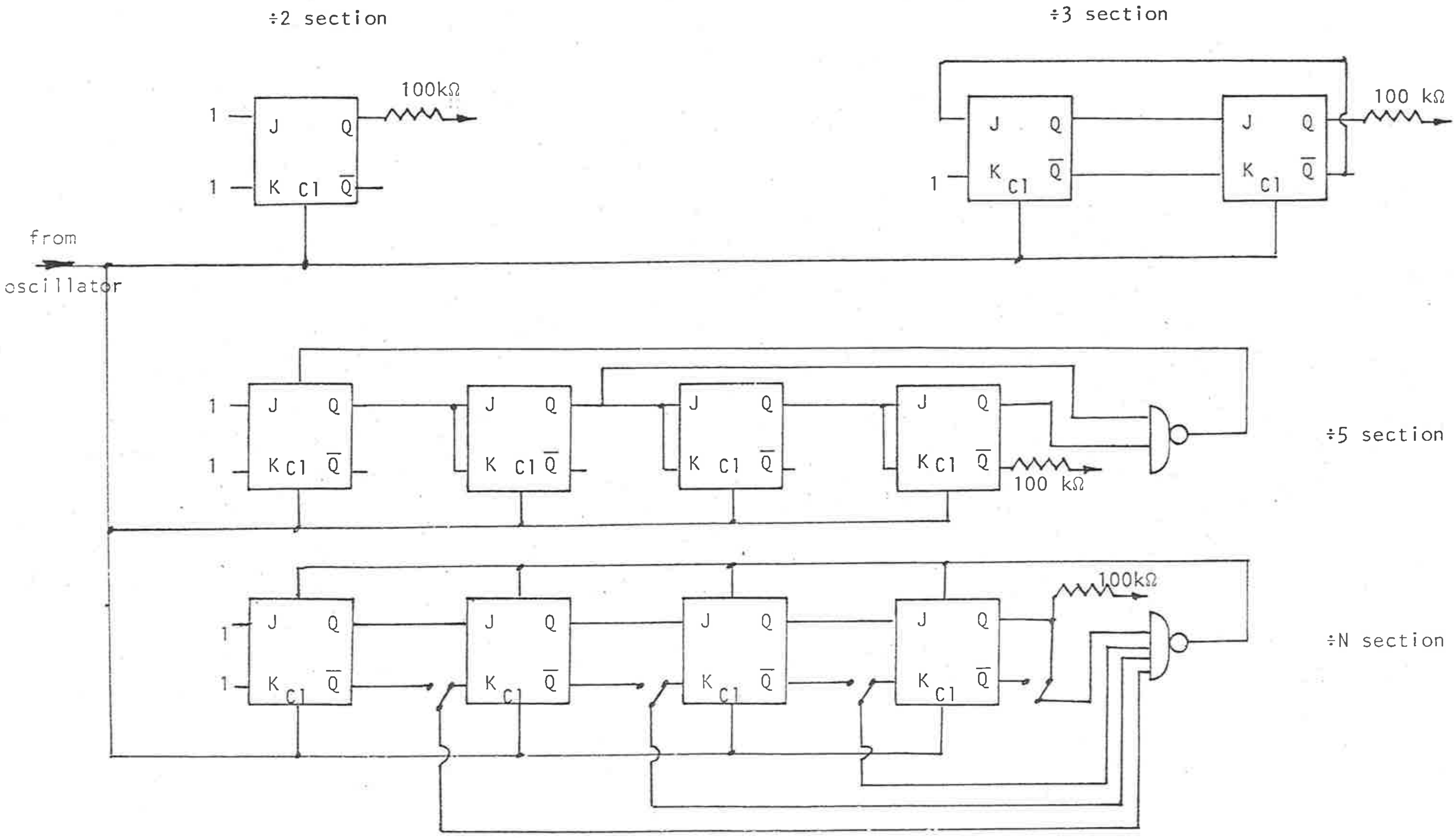


DIAGRAM 3.4.D. Logic section of the digital division oscillator.

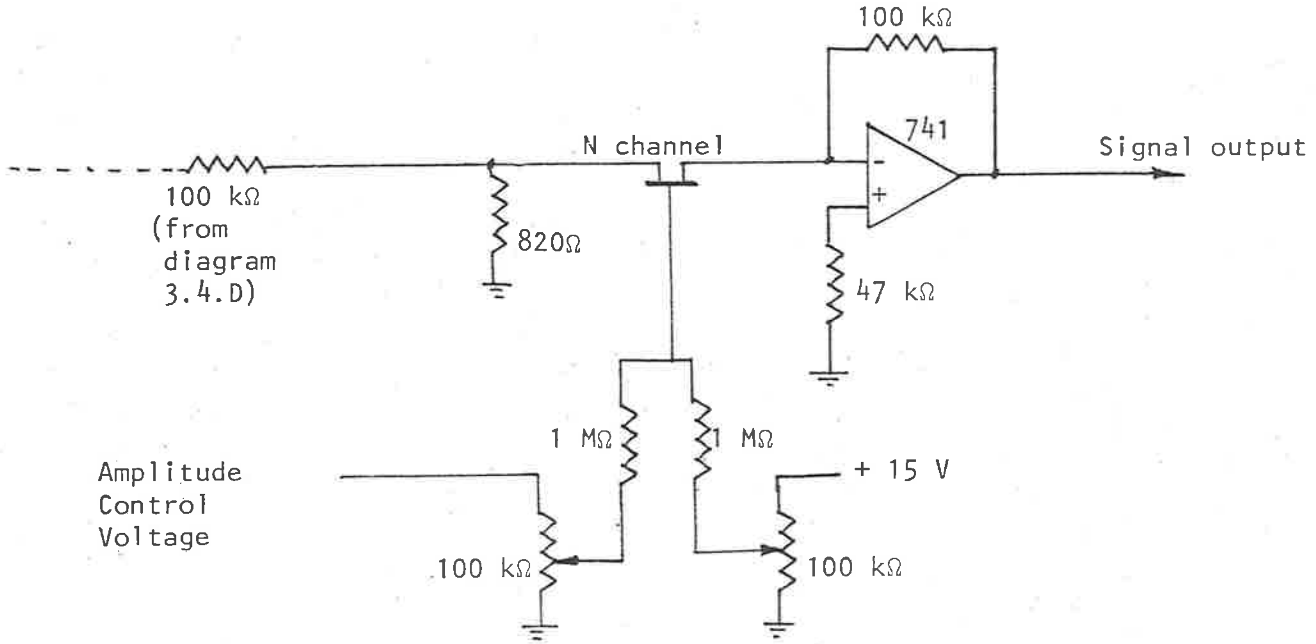


DIAGRAM 3.4.E. Analog section of the Digital Division Oscillator.

3.5 White Noise Generator

The White Noise Generator is the only non-periodic sound source. The random signal is obtained by amplifying the noise which is present across a zener diode which is conducting current in the reverse direction. This unit provides one signal output voltage with an amplitude which is voltage controlled, and a separate control voltage output.

The signal output is filtered. The cutoff frequencies of this filter are controlled by potentiometers which must be set manually. Voltage controlled filtering can be obtained using the filter unit described in section 4.3. The advantage of providing a relatively simple manual filter is that is allows the voltage controlled filter to be used for other functions in most cases. This manual filter consists of a high pass Sallen and Key filter, and also a low pass Sallen and Key filter which are both second order. These can be cascaded or used in parallel so that high pass, low pass, band pass, and band stop characteristics are available.

The block diagram of the White Noise Generator is given in diagram 3.5.A, and the circuit diagram is given in diagram 3.5.B. and diagram 3.5.C.

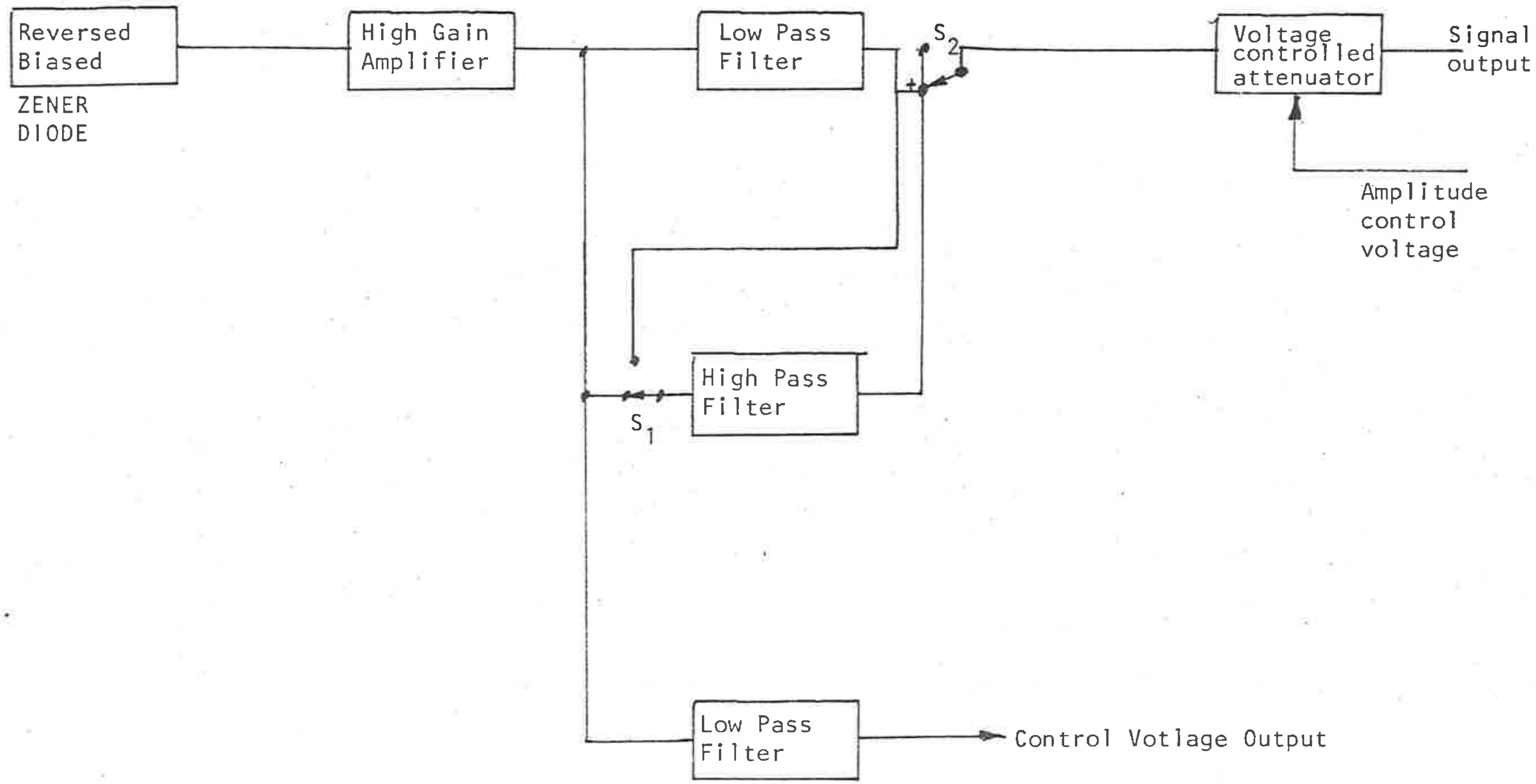


DIAGRAM 3.5.A. Block diagram of the white noise generator

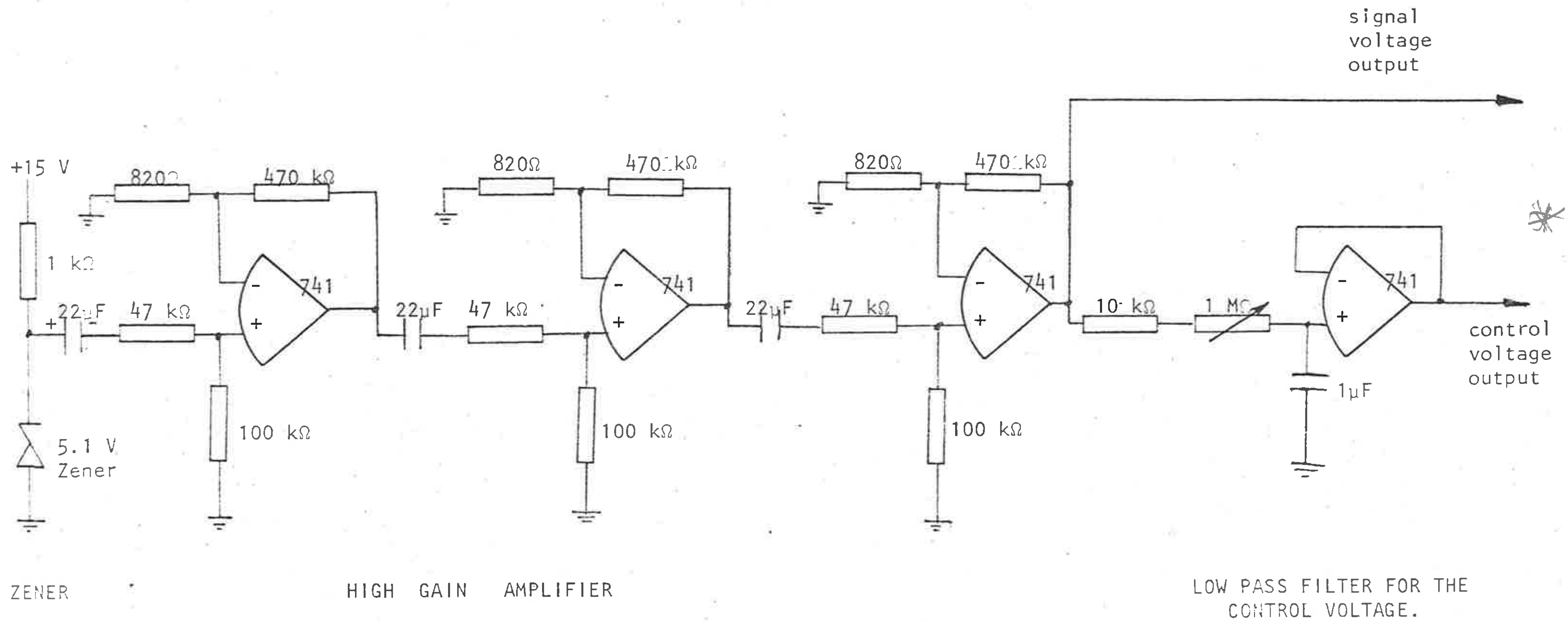


DIAGRAM 3.5.B. Circuit of the noise generator section of the White Noise Generator.

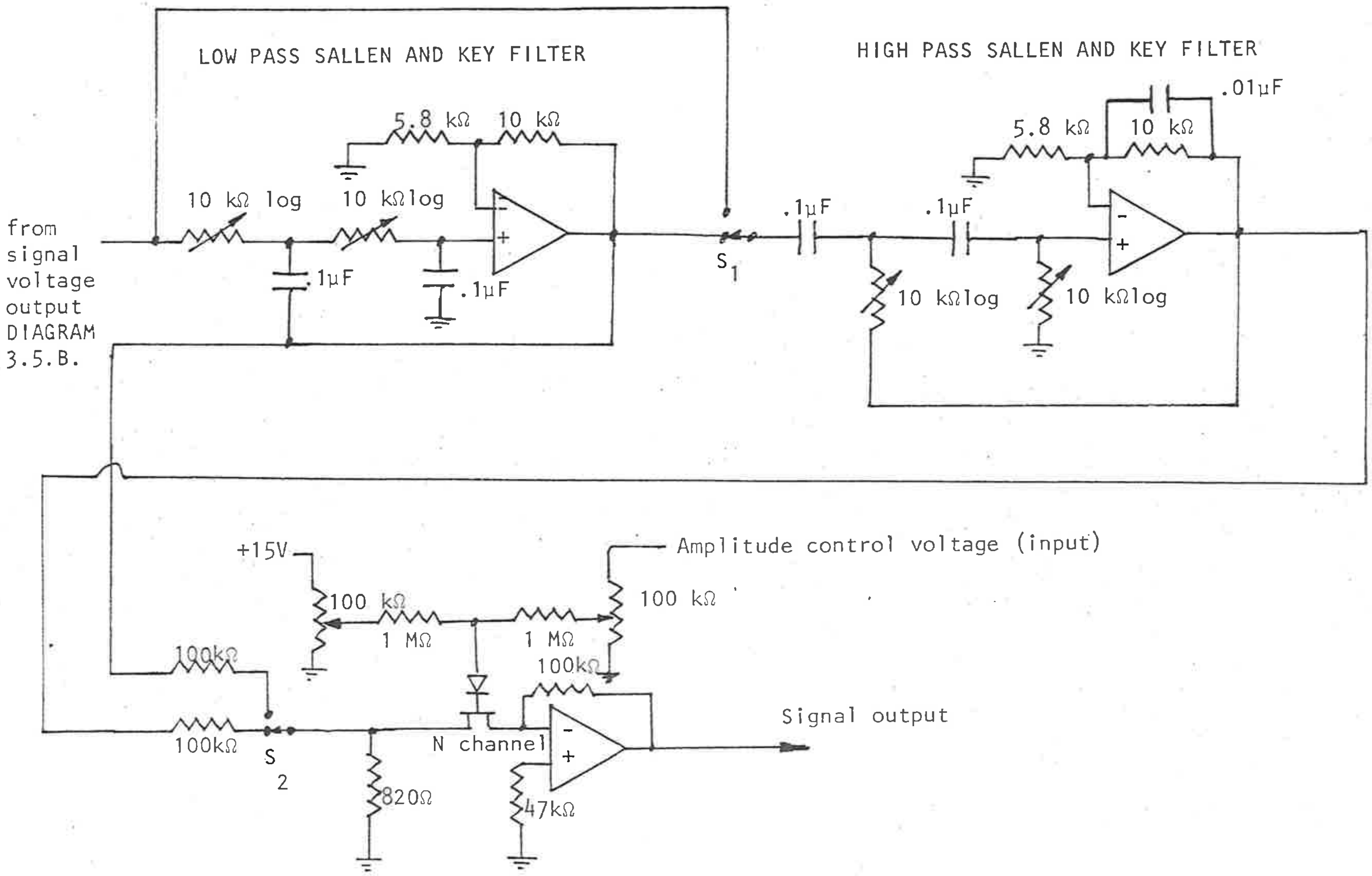


DIAGRAM 3.5.C. Filter section of the White Noise Generator.

NOTES FOR CHAPTER 3

1. When the frequency control voltage contains some signal components complex frequency modulation spectra result. This effect is discussed from the musician's point of view in reference 5.

CHAPTER 4 SOUND MANIPULATORS

4.1 Importance of Sound Manipulators

Sound manipulators are characterized by having signal inputs as well as signal outputs. These units are an important aspect of a music synthesizer. It is desired to provide a wide variety of sounds for the musician. The sound sources and sound manipulators are designed so that each combination gives a distinct type of sound. The number of combinations of these units exceeds the total number of units, and in this way the sound manipulators extend the variety of sounds available from a given amount of equipment.

The proportion of the resources which should be devoted to sound manipulators can be estimated by assuming the cost of each unit is approximately the same, and each combination of units gives a distinct type of sound. The sound produced when sound manipulator SM_1 is followed by sound manipulator SM_2 usually differ from the case when the order of the sound manipulators is reversed and SM_2 is followed by SM_1. In this case the number of distinct types of sound produced by selecting sound manipulators is given by the number of their permutations. For example the number of permutations of 2 units when 9 are available is 9 x 8 = 72. In general the number of permutations of n objects chosen from m distinct objects is m!/n!, and is written P_n^m.

For linear sound manipulators the sounds produced do not depend upon the order of sound manipulators, and the number of distinct sound types produced is given by the number of combinations of n objects chosen from m objects. In this case this number is $C_n^m = m!/(n!(m-n)!)$.

A table and a corresponding graph is given to show the importance of sound manipulators, and to assess the proportion of resources devoted to sound manipulators to the total resources of the synthesizer. The table and graph show the number of permuations available. This is the best estimate of the number of distinct types of sound because the use of any nonlinear sound manipulator means that the sequence of the sound manipulators is important.

TABLE 4.1.A

Number of Sound Source Units N_{ss}	Number of Sound Manipulator Units N_{sm}	Available permuations of sound Manipulators with; no: manipulators	one:	two;	three;
10	0	10	0	0	0
9	1	9	9	0	0
8	2	8	16	16	0
7	3	7	21	42	42
6	4	6	24	72	144
5	5	5	25	100	300
4	6	4	24	120	480
3	7	3	21	126	630
2	8	2	16	112	672
1	9	1	9	72	504
0	10	0	0	0	0

Permutations of sound sources and sound manipulators for various ratios of sound sources to sound manipulators.

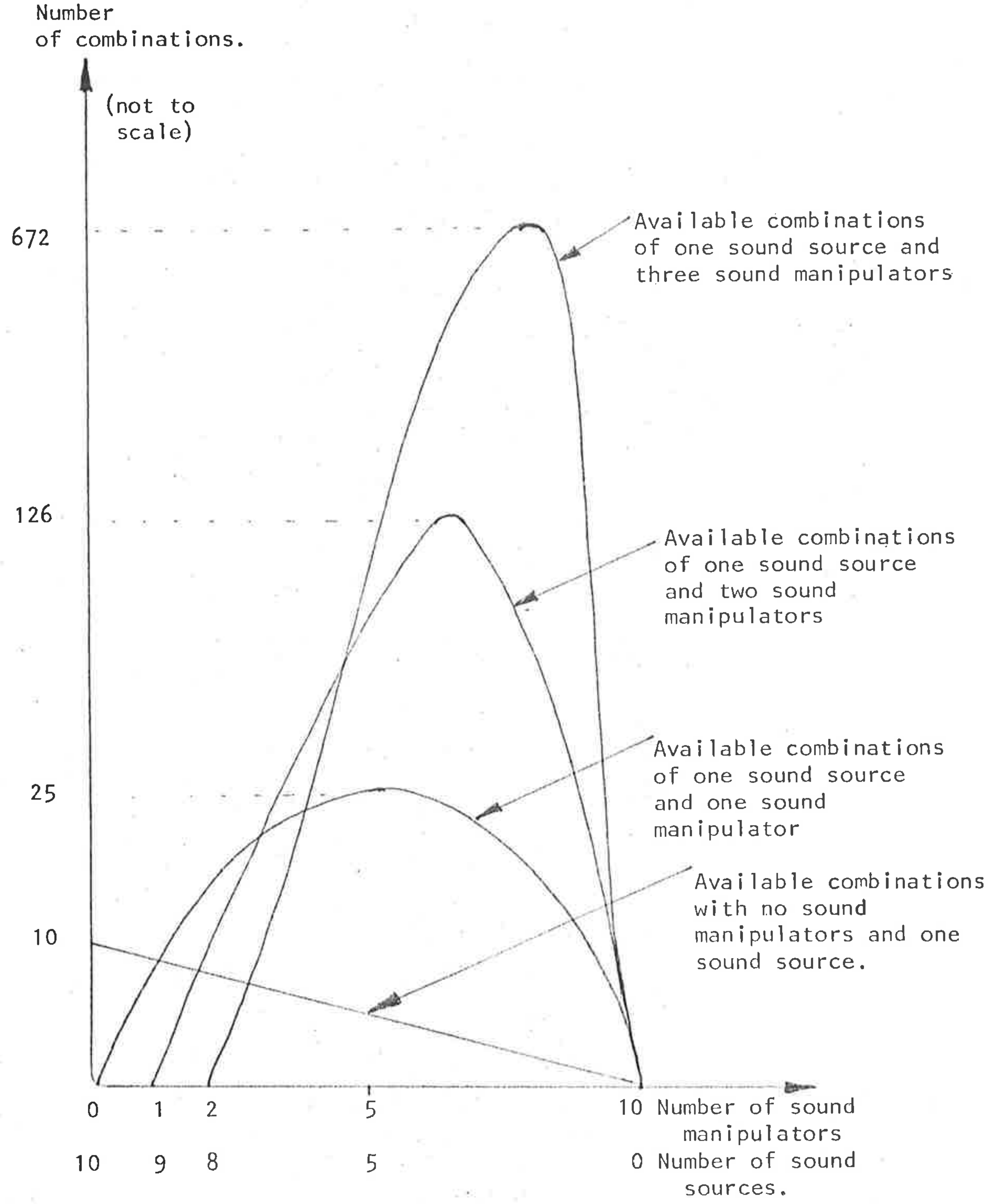


DIAGRAM 4.1.A. The number of available combinations of sound manipulators and sound sources for various ratios of the number of sources to manipulators assuming a total of ten units for the synthesizer.

If combinations of one sound source and one sound manipulator are considered the number of sound manipulators should equal the number of sound sources to give the maximum number of permutations. When combinations of one sound source and two or more sound manipulators are considered, the number of available combinations increases considerably, and the number of sound manipulators should actually exceed the number of sound sources to give the maximum number of combinations. In fact, according to the assumptions, the greatest number of available sounds occurs when there is only one sound source and the remaining resources are devoted to nine sound manipulators. This is because the number of arrangements of nine sound manipulators is so large. However there are complementary factors to be considered. It is convenient to produce more than one independent sound at a time. This requires more than one sound source. Also it is possible to use more than one sound source to produce a given sound. The more units introduced involved producing a sound,the more difficult is the operation of the synthesizer. Completely different sounds are available from different sound sources and would increase the range of available sounds, whereas sound manipulators give some similarity in the sound produced.

It was decided to use six sound manipulators compared with four sound sources.

4.2 Application of the Design Approach

4.2.1 Operator Control

The relationship between the input and output of most of the sound manipulators can be controlled by a voltage. In this way the operator can use a control unit to produce the sound. For example the passband frequency of the filter can be determined by the position of a lever. This is to allow real time control over the sound produced.

4.2.2 Diversity

The units are designed to give very different effects on the sound. The Harmonic Generator and the Frequency Divider add frequency components which are above and predominantly below the input frequency respectively. The multiplier adds frequency components which depend upon two inputs. The Filter and Reverberation units do not add frequency components, but change those present in very different ways. There is provision for position control of the signal by determining the ratio of the amplitudes in two channels, and this is voltage controlled.

Each unit is designed to have outputs which are capable of a wide range of sound. The filter has high, low and band pass outputs; the output of the Frequency Divider preserves components of the input waveform rather than the uniform square wave of logic circuitry which is used for the frequency division.

4.2.3 Predictability

When a synthesizer is used by a musician it is necessary for him to be able to assess its scope and its limitations, and be able to achieve an intended result. The sound manipulators are designed to give predictible transformations. The effect of a filter on a sound can be predicted, and this allows the musician to systemmatically use this facility. A number of interesting electronic effects are almost impossible to set up again. These effects may depend upon some critical triggering conditions, or perturbations to a complex oscillatory condition. In each of these sound manipulators a specific input to output relationship has been achieved.

4.3 Voltage Controlled Filter

A filter is a very useful sound manipulator in a synthesizer. It is a powerful technique for modifying sounds, and the effect can be readily appreciated by the musician. The objective is to design a versatile filter which can be controlled by the musician. The filter for the synthesizer has high pass, low pass and band pass outputs, and its transfer characteristics are voltage controlled.

4.3.1 Second Order Filter

A filter based on the biquad principle[1] was used in this application. A simplified circuit of this second order filter and the corresponding block diagram are given in diagrams 4.3.A. and 4.3.B. In general this filter has a transfer function of 2 poles and 2 or less zeros.[2]

$$T(s) = \frac{a_2 s^2 + a_1 s + a_o}{b_2 s^2 + b_1 s + b_o}$$

$$\text{where } a_2, a_1, a_o \geq 0$$
$$b_2, b_1, b_o > 0$$

Diagram 4.3.C shows the high pass, low pass and band pass transfer functions, the corresponding s plane representations and the Bode magnitude plots. In general the zeros can be on the real axis or complex conjugates. This circuit is chosen to locate the zeros at the origin because this gives a more suitable magnitude response in the bandpass and high pass filters.

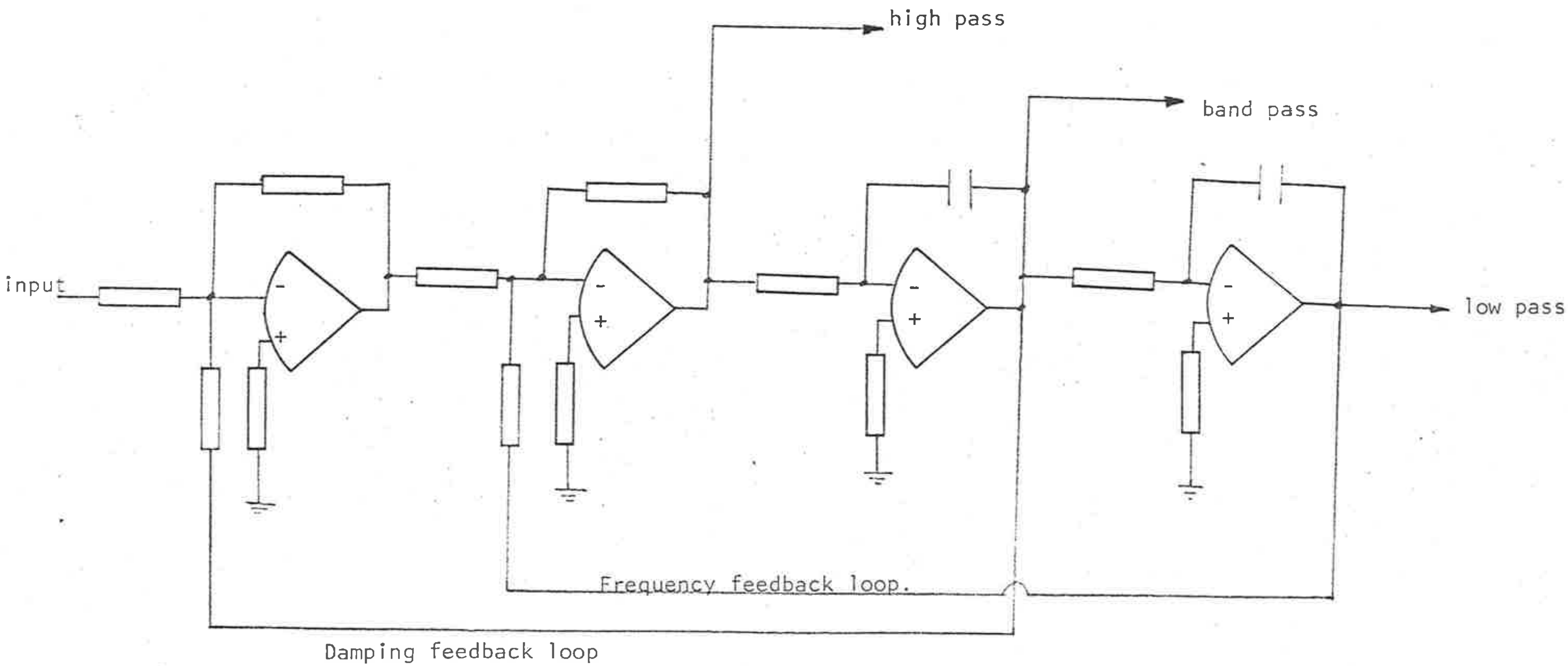


DIAGRAM 4.3.A. Simplified circuit of the second order biquad filter.

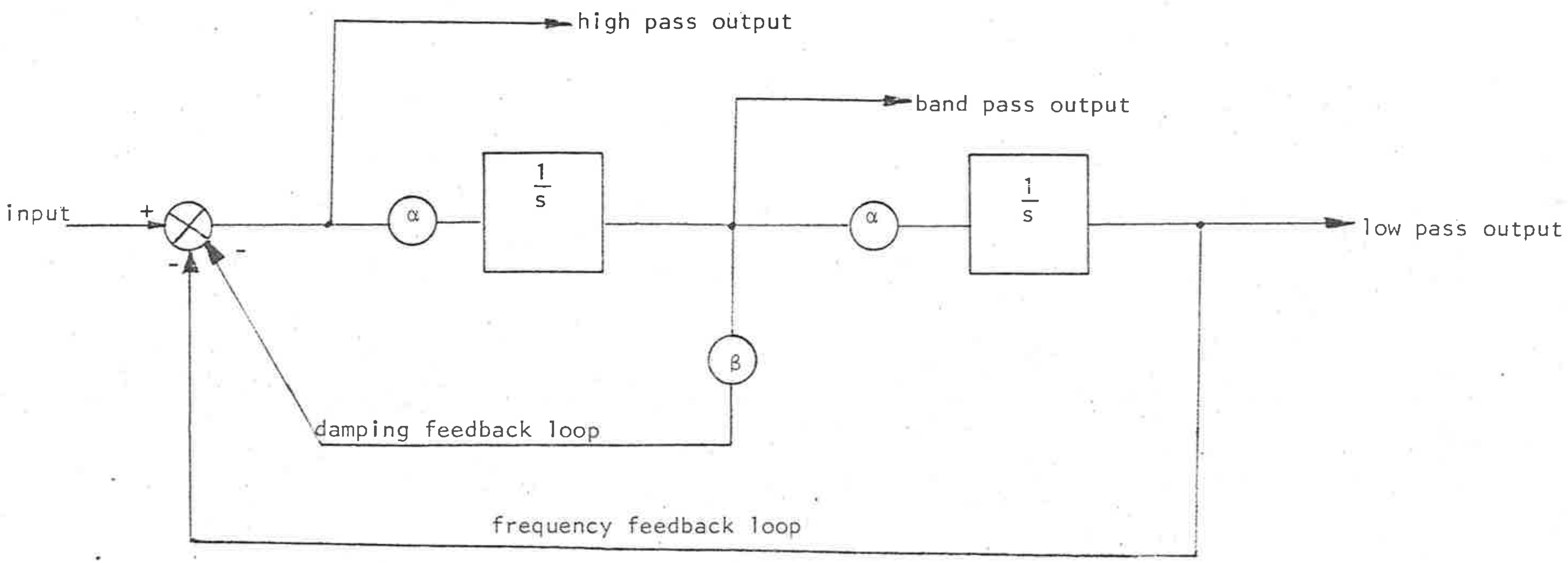


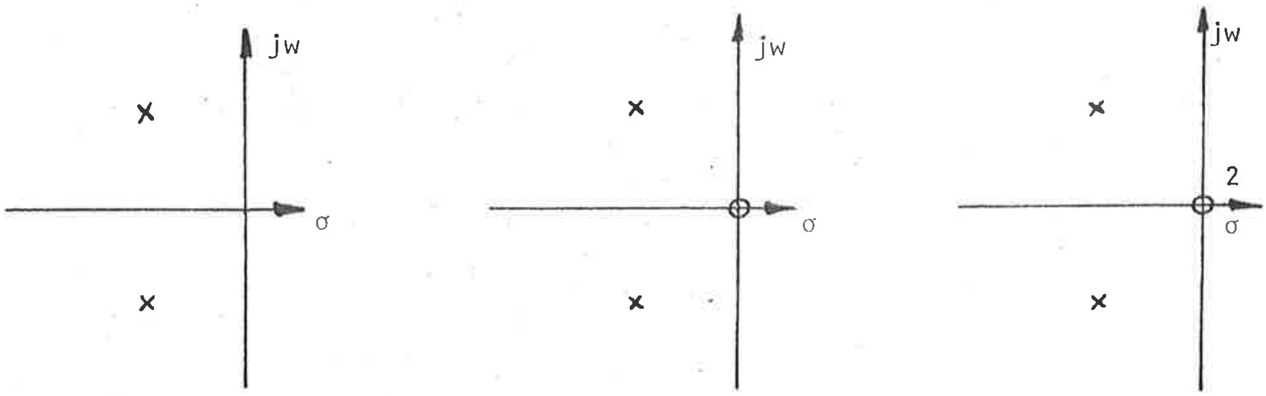
DIAGRAM 4.3.B. Control system representation of the second order filter.

Transfer function

$$T(s) = \frac{\alpha^2}{s^2 + s\alpha\beta + \alpha^2} \qquad T(s) = \frac{s\alpha}{s^2 + s\alpha\beta + \alpha^2} \qquad T(s) = \frac{s^2}{s^2 + s\alpha\beta + \alpha^2}$$

LOW PASS   BAND PASS   HIGH PASS

S plane representation



Bode magnitude

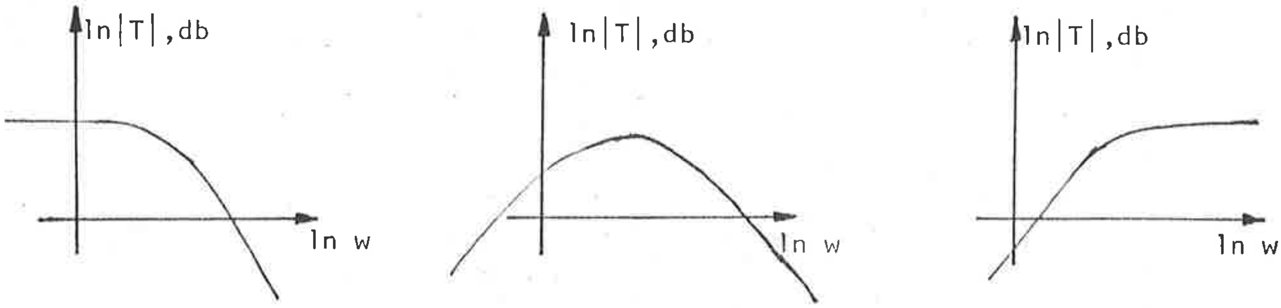


DIAGRAM 4.3.C. High pass, band pass and low pass transfer functions s plane representations and Bode magnitude plots.

4.3.2 Voltage Control

Voltage control is used to change the position of the poles. This movement must suit the musician's requirements. Diagram 4.3.D gives suitable and unsuitable loci. The suitable locus is such that the damping factor is unchanged as the cutoff frequency is changed. There is a second control voltage for the damping factor which can be controlled independently of the cutoff frequency.

The transfer functions for the filter of diagram 4.3.A can be found using block diagram manipulation. Diagram 4.3.E is used to show the low pass transfer function is,

$$T(s) = \frac{\alpha^2}{s^2 + s\alpha\beta + \alpha^2}$$

where α and β are gain parameters on the block diagram of 4.3.D

The bandpass output is present at the input to the integrator which has the low pass output, $C(s)$. It follows that the bandpass output is $sC(s)/\alpha$. Similarly the high pass output is $s^2C(s)/\alpha^2$. The corresponding transfer functions are;

$$\text{Bandpass} \quad sT(s) = \frac{s\alpha}{s^2 + s\alpha\beta + \alpha^2}$$

$$\text{High pass} \quad s^2T(s) = \frac{s^2}{s^2 + s\alpha\beta + \alpha^2}$$

These polynomial expressions giving the poles and zeros of the transfer functions illustrate that the desired loci of the poles will be achieved if the term 'α' is voltage controlled. This corresponds to altering the gains of both the integrators and results in a frequency shift without a change in the Q of the filter. Similarly the damping

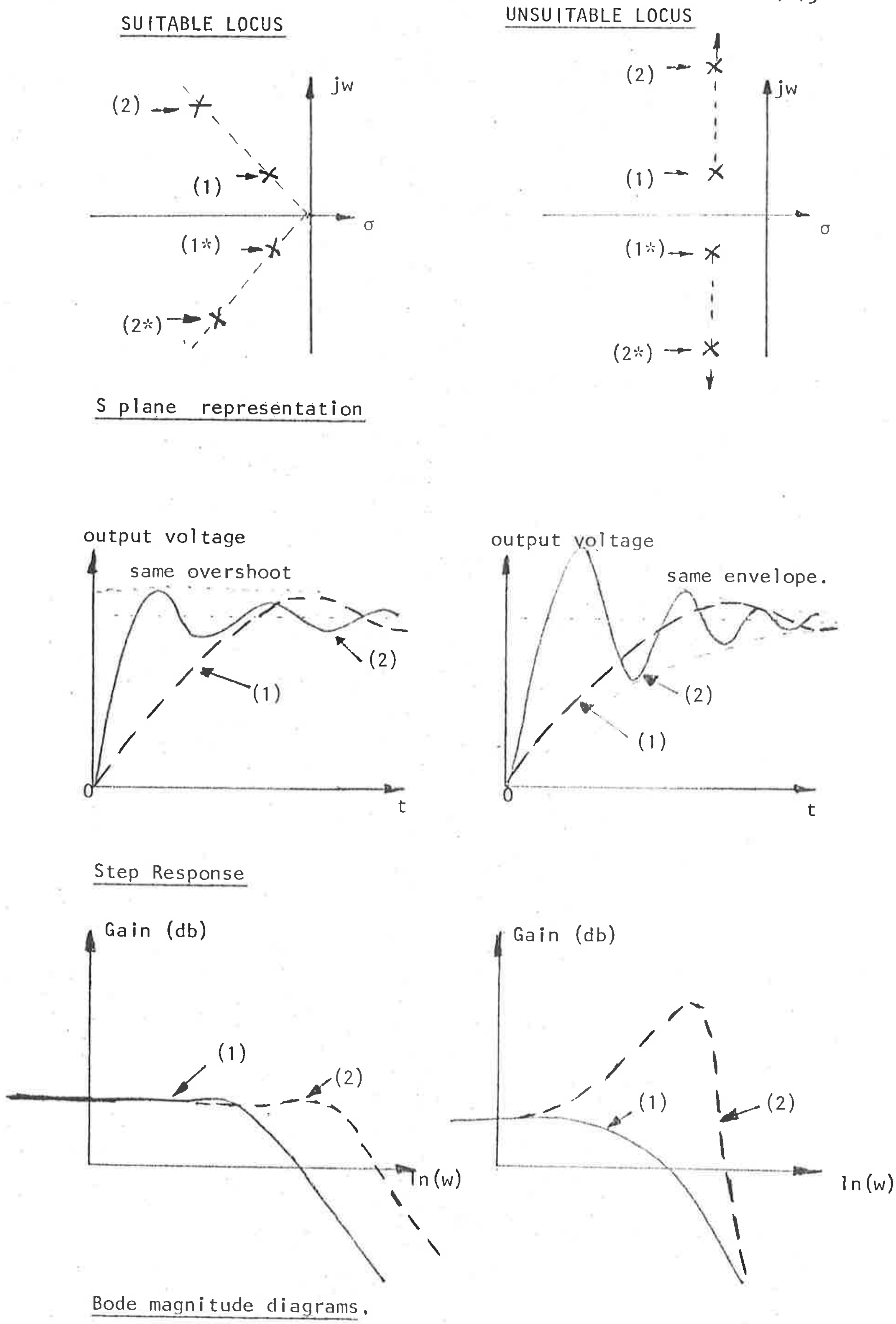
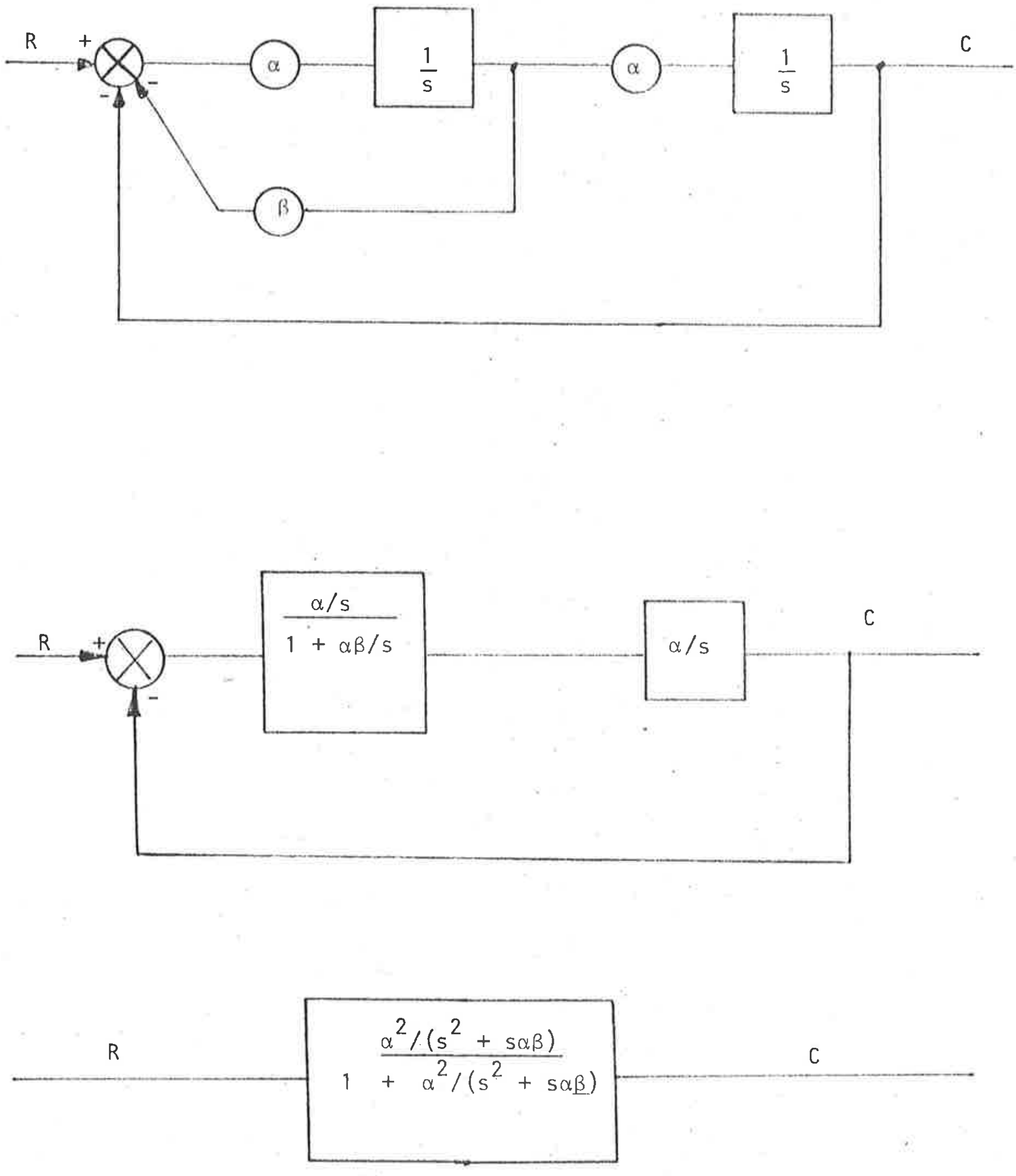


DIAGRAM 4.3.D. Suitable and unsuitable loci for the poles of the low pass filter as the frequency control votlage is changed.



$$T(s) = \frac{\alpha^2}{s^2 + s\alpha\beta + \alpha^2} = \frac{w_n^2}{s^2 + 2s\xi w_n + w_n^2} \qquad \text{where } \alpha = w_n \text{ and } 2\beta = \xi$$

DIAGRAM 4.3.E. Obtaining the transfer function of the filter using control system block diagram manipulation.

factor or Q depends upon the coefficient ' β', only.

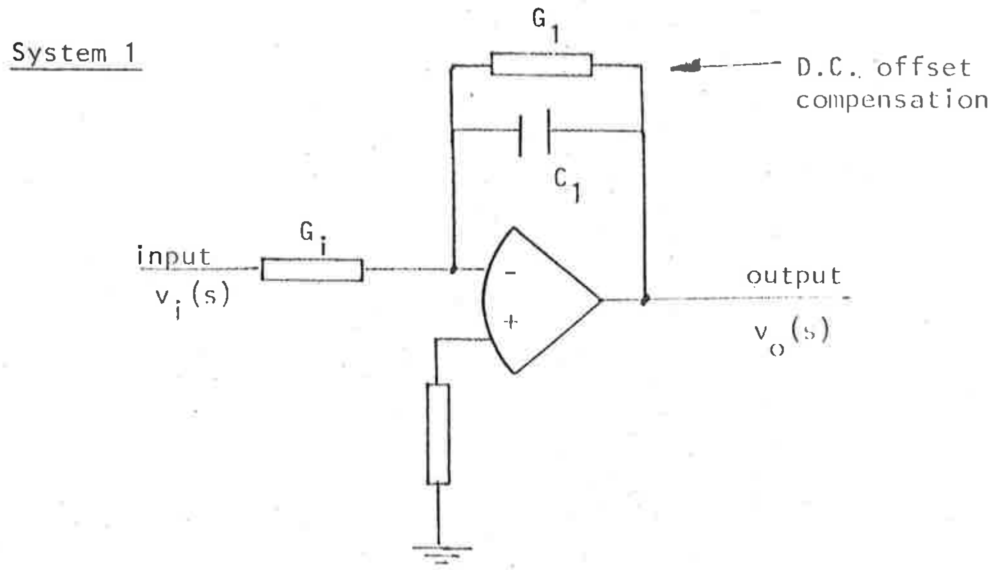
4.3.3 Analog Multipliers - Feedthrough

The coefficients were voltage controlled using analog multipliers. The desired frequency range is 10^3:1, (5 Hz to 5 kHz), and the corresponding ratio of the coefficient α is also 10^3:1. The feedthrough of the XR 2208 analog multiplier is .1 V, which is 10^{-3} times the maximum voltage available at the output of the multiplier. This means the feedthrough is unacceptable at low frequencies. Two analog multipliers can be used in cascade to extend the frequency range.

4.3.4 Analog Multipliers - Offset

The D.C. offset voltage at the output of the analog multiplier is as high as .14 V, and this is amplified by a factor of 10 by the buffer associated with the analog multiplier to give an uncompensated D.C. output offset of 1.4 V. In practice the offset adjustment can keep this offset to within ±.1 V over the commercial temperature range of 0°C to 70°C. The large D.C. gain of the integrator (ideally infinite) means that the D.C. stabilization of the filter circuit is achieved by the action of the main feedback loop of the filter, and no subsidiary loop is required. For small values of the gain term α a small voltage at the output of the multiplier causes an unacceptably large output of the integrator.

Diagram 4.3.F shows a suitable D.C. stabilization circuit for the integrator. This bypass arrangement allows a small sensitivity in the output to D.C. offsets present at the input, with a smaller change to the low frequency response of the integrator than that introduced by a simple resistive shunt.



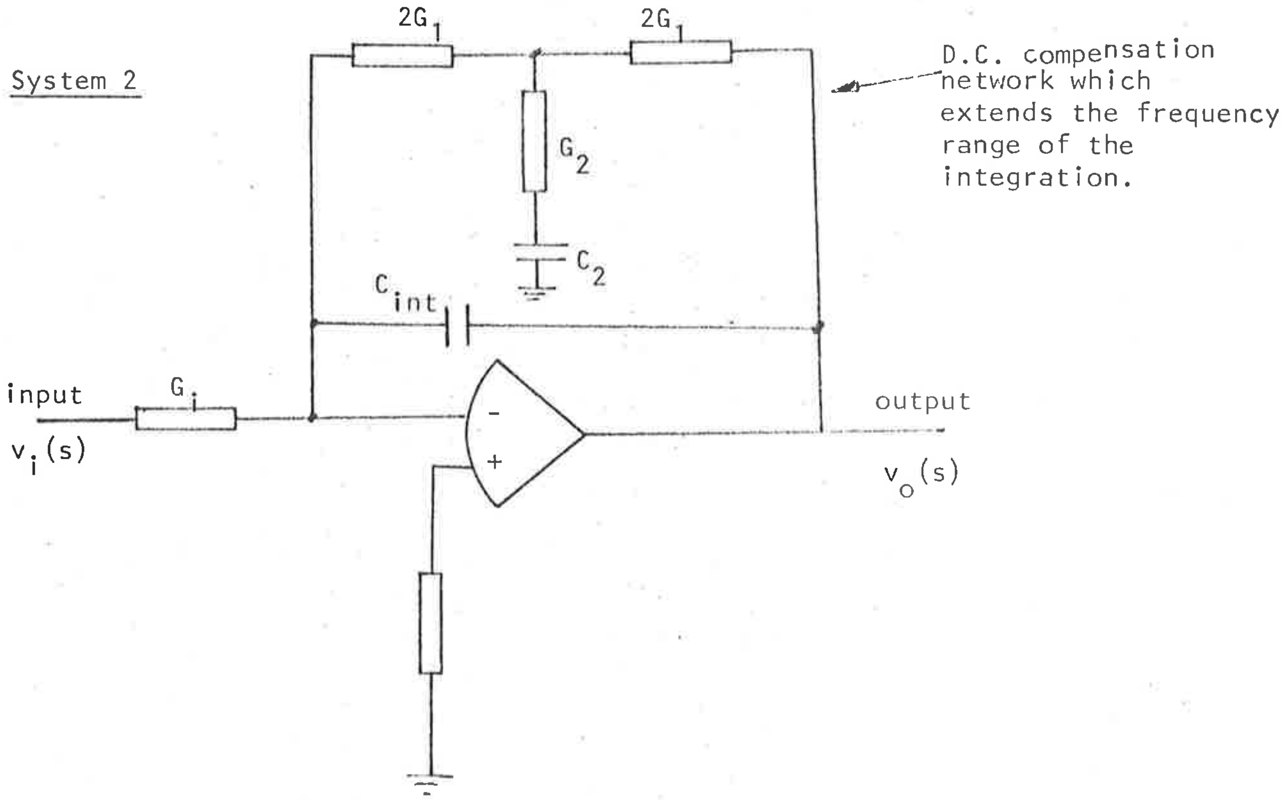


DIAGRAM 4.3.F. Two techniques for compensation for the D.C. offset voltage into a partial integrator. For a given D.C. gain for the stage, the second system has an extended frequency range of integration (see Diagram 4.3.G.)

The transfer function for the simple compensation system, 1, is;

$$T_1(s) = \frac{v_o}{v_i} = -\frac{G_i}{sC_1 + G_1}$$

The general transfer function for the compensation system which gives the extended frequency range is;

$$T_2(s) = -\frac{G_i}{sC_1 + 2G_1 \dfrac{2G_1}{4G_1 + \dfrac{G_2 sC_2}{G_2 + sC_2}}}$$

$$= -\frac{G_i(s(4C_2G_1 + 4C_2G_2) + 4G_1G_2)}{2s^2C_1C_2(4G_1+G_2) + s(4G_1{}^2C_2 + 4G_1G_2C_1) + 4G_1{}^2G_2}$$

but $C_2 \gg C_1$ and $G_2 \gg G_1$ by design.

Hence $$T_2(s) \simeq -\frac{G_i(2G_1 + sC_2)}{s^2C_1C_2 + s\left[\dfrac{4G_1{}^2C_2}{G_2} + 4G_1C_1\right] + 4G_1{}^2}$$

The transfer function $T_1(s)$ gives a pole at $-G_1/C_1$. The transfer function $T_2(s)$ has the same D.C. gain because the total feedback conductance is G_1 with both systems. However the transfer function has two poles and a zero. By suitable design the poles can be made coincident on the real axis. For this case the determinant of the quadratic equation in s (the characteristic equation) is zero.

Hence $$\frac{4G_1{}^2C_2}{G_2} + 4G_1C_1 = (4G_1{}^2C_1C_2)^2$$

For $C_2 \gg C_1$ it follows that $$\frac{G_1}{G_2} \simeq \tfrac{1}{2}\sqrt{\frac{C_1}{C_2}}$$

When the poles for system 2 are located together on the real axis they are at $-2G_1\sqrt{\frac{C_1}{C_2}}$ and the zero is at $-2G/C_2$. These locations involve considerably longer timeconstants than the pole of the first system as is shown on Diagram 4.3.G. This diagram gives the s plane location of the singularities and the corresponding frequency responses of the two systems.

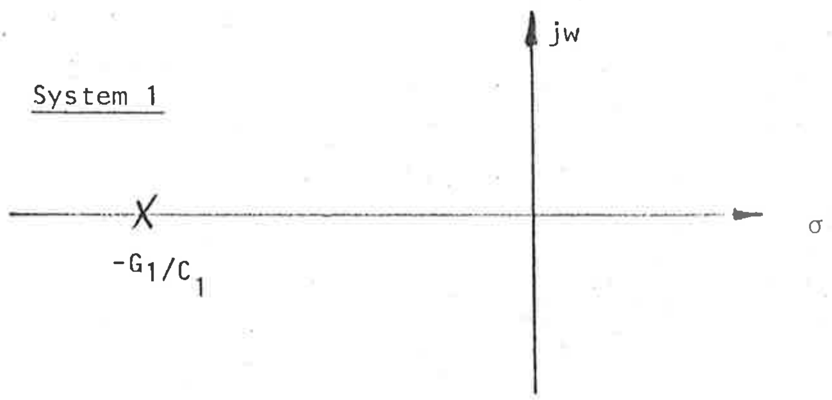
4.3.5. Control Voltage to Frequency Relationship

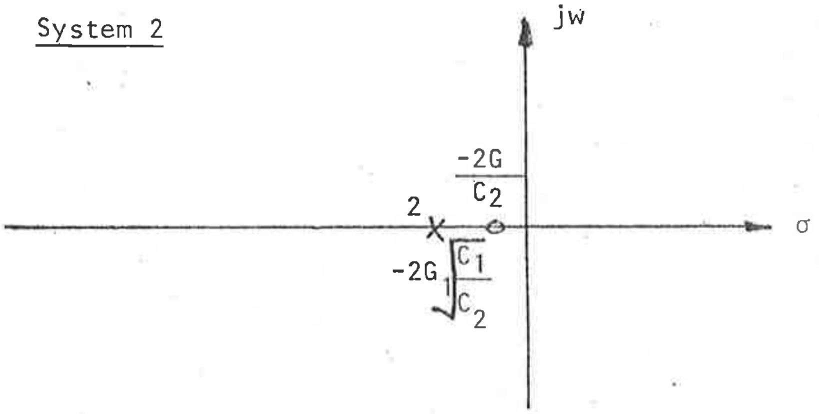
It is necessary to design a nonlinear resistive element to give a logarithmic relationship between the frequency control voltage v_c and the corresponding cutoff frequency of the filter w_o. It is required that;

$$w_o = k_1 e^{k_2 v_c}$$

where k_1 and k_2 are suitable constants.

From the transfer function relationships derived in diagram 4.3.E. the cutoff frequency w_o is proportional fo the coefficient α, and in turn α is proportional to the square ot the control voltage applied to the multipliers because there are two multipliers in cascade. Hence; $w_o = k_3 v_m^2$, where v_m is the voltage applied to the multipliers which is logarithmically related to the control voltage v_c, and k_3 is a contant which depends upon the integration rate of the integrators in the filter. The optimum operating range of the voltage to the multipliers is when this range includes the maximum allowable voltage level of +10 V. It follows that the minimum voltage level for a frequency range of 10^3:1 is $10 : \sqrt{10^3}$ which is approximately .3 V. Therefore it is necessary to design a nonlinear resistive element which will take the control voltage range





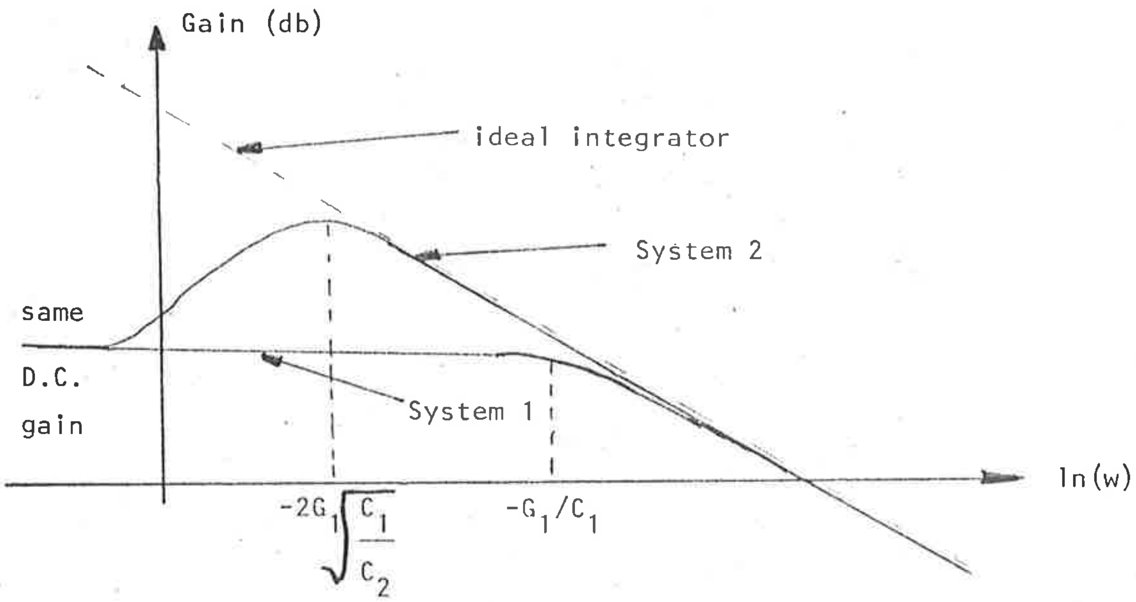


DIAGRAM 4.3.G. Comparison of two techniques for compensating the integrator for input voltage offsets. The D.C. gains of the two systems are equal, yet the frequency at which System 2 still integrates is lower than System 1.

input of +5 V to -5V and convert it to a logarithmically dependent voltage with corresponding extrema of 10 V and .3 V respectively. The general form of the required relationship is;

$$V_m + k_4 = k_5 \log_{10} V_c$$

where k_4 and k_5 are constants which satisfy the following pairs of voltage levels.

$$V_c = -5 \quad V_m = .3 \quad \text{and} \quad V_c = +5 \quad V_m = 10$$

The circuit of Diagram 4.3.H. gives a nonlinear resistor as the input resistor to an operational amplifier. Seven segments are used. The values of the resistors in these segments can be calculated knowing that there will be a voltage drop of .7 V across each diode during conduction. Table 4.3.A. shows the desired output voltages of this circuit, V_m, which correspond to the limits of each of these seven linear segments. The required changes to the output voltage across each section can then be calculated. The required incremental resistances of the nonlinear resistor can then be found knowing that, for an inverting operational amplifier;

$$|\Delta V_{in}/\Delta R_{in}| = |\Delta V_{out}/\Delta R_{fb}|$$

where R_{in} is the input resistor to the operational amplifier, and R_{fb} is the resistor in the feedback path. In this case ΔV is .7 V and ΔR_{fb} is constant and equal to 10 kΩ by design choice. Hence;

$$\Delta R_{in} = \frac{|\Delta V_{in}|}{|\Delta V_{out}|} \cdot R_{fb} = .7 \times 10\ \text{k}\Omega\ /\ \Delta V_{out}.$$

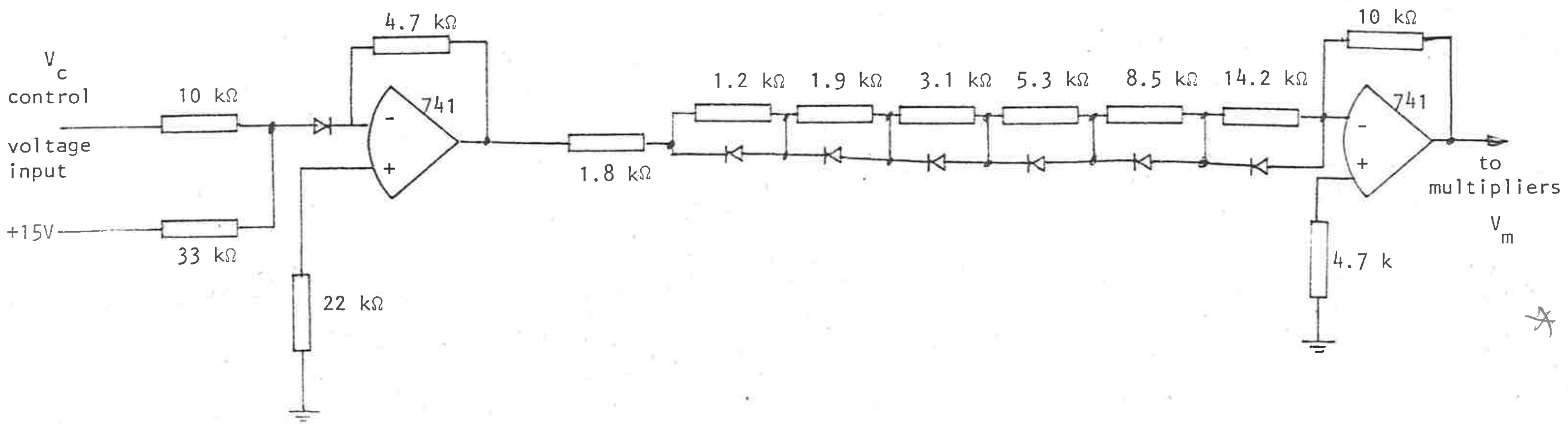


DIAGRAM 4.3.H. Circuit to give the nonlinear transformation between the control voltage V_c and the voltage fed to the multipliers V_m, so that a logarithmic control voltage to frequency relationship is obtained.

Segment Conducting	Desired Output Voltage	Change in output voltage across section	Required incremental resistance (kΩ)	Additional resistance of section (kΩ)
1	.30	.195	35.9	14.2
2	.495	.322	21.7	8.5
3	.817	.531	13.2	5.3
4	1.35	.882	7.94	3.1
5	2.23	1.44	4.85	1.92
6	3.67	2.39	2.93	1.15
7	6.06	3.94	1.78	1.8 (total)
	10.0			

TABLE 4.3.A. Calculation of the resistors required in the nonlinear resistive element of the circuit of diagram 4.3.H.

Then the additional resistance for each section is tabulated.

The input to the nonlinear resistor must be from 0 to 5 V, and the control voltage range is from -5 to +5 V. The required offset and magnitude change is provided by the first integrated circuit of Diagram 4.3.H.

4.3.6 Damping Factor Control

The damping factor is proportional to the coefficient β in the transfer function derived in Diagram 4.3.E. When $\beta = 0$ there is no damping, though the system tends to be stable because the integrators have D.C. bypass resistors which provide a stabilizing effect. The range of damping factors of 0 to 1 gives a suitable variety of filter characteristics. Diagram 4.3.I. gives the loci of the poles as the damping factor is altered, and the corresponding Bode magnitude plots of the low pass filter.

The circuit of the Filter is given in Diagram 4.3.J.

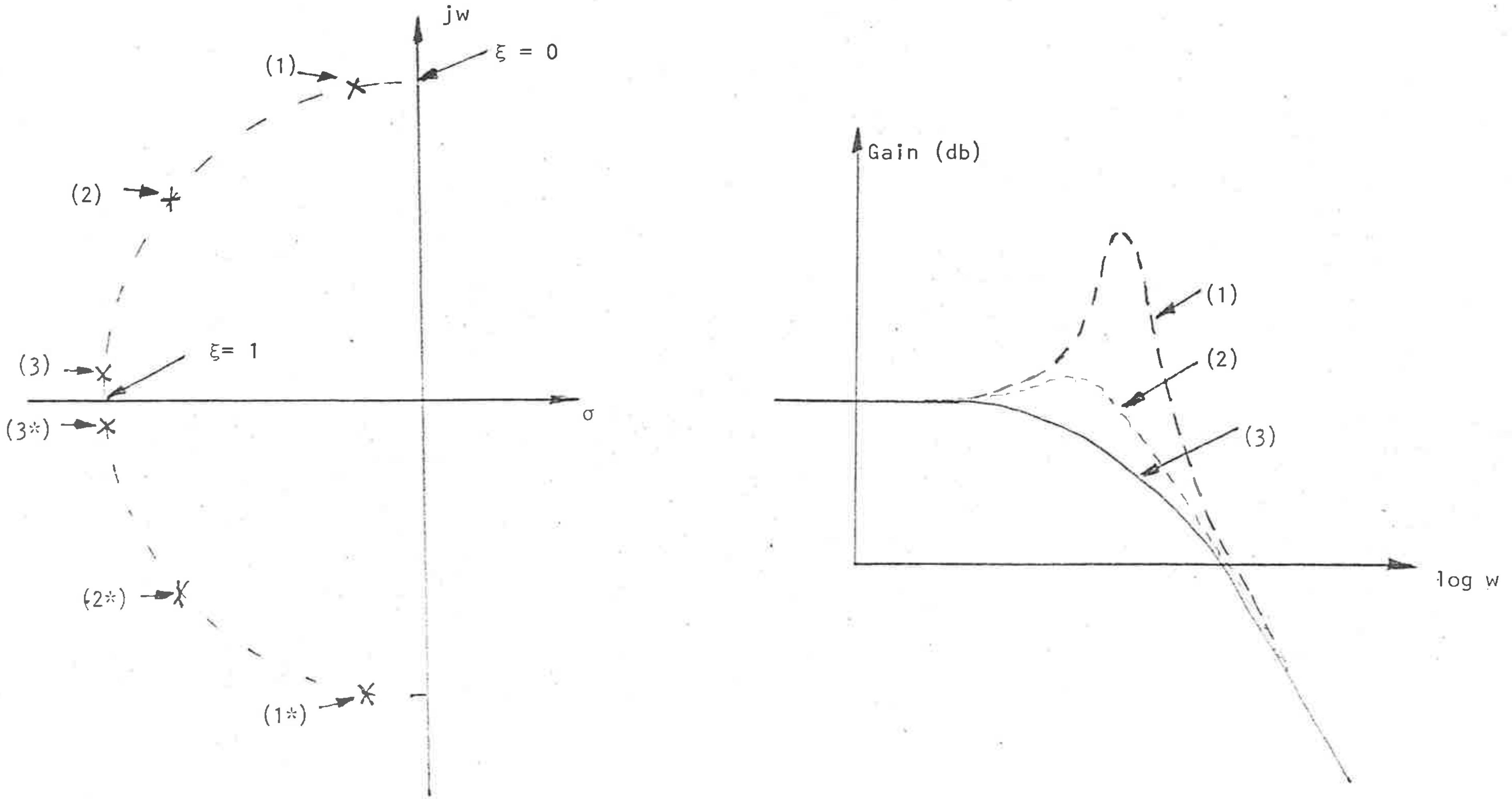


DIAGRAM 4.3.1. Locus of the poles of the low pass filter as the damping factor is altered, with the corresponding Bode magnitude plot.

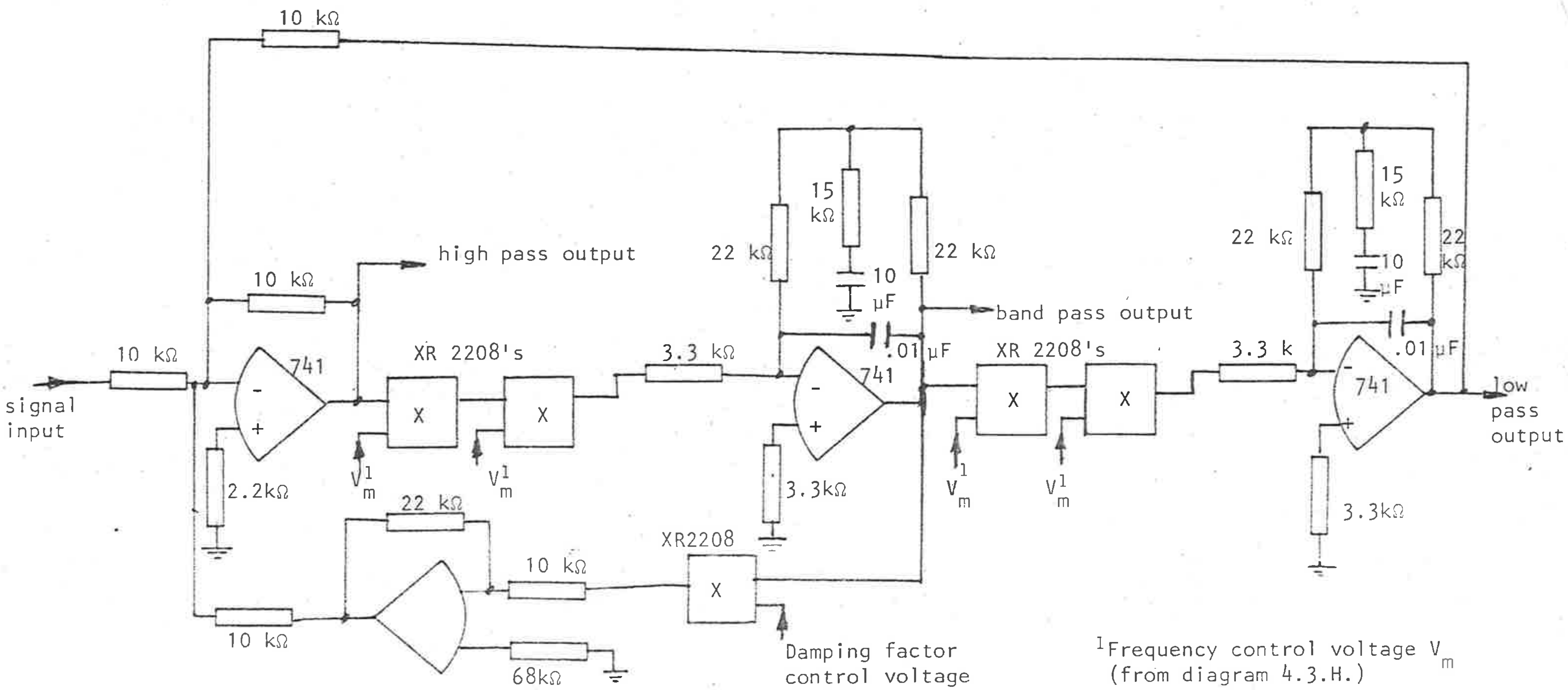


DIAGRAM 4.3.J. Circuit for the voltage controlled filter.

4.4 Multiplier Resigned *

Signal multiplication is a very useful technique for sound manipulation. In engineering terms the output frequencies are given by the sum and the difference of the frequencies present in the signal input. In practice signal multiplication is a unique way of combining two signals in such a way that characteristics of both input sounds are present, yet a radically different sound is produced. The multiplier can also be used with control voltages as well as signal voltages.

Commonly available integrated circuits are suitable for this application. The XR 2208 has specifications of ±3% linearity and a bandwidth of 8 MHz at 3 db. This is quite adequate.

These integrated circuit multipliers were not available when the requirement for a multiplier first arose, and s suitable four quadrant multiplier was designed. This design uses switching techniques, and may still have applications when suitably refined. The principle is capable of excellent linearity because of the high switching speeds which are now available, This linearity is required when detecting signals of low level in the presence of large noise components.

Two multiplier circuits using the principle of time-division[3] are given. The first, which uses only three linear operational amplifiers to achieve four quadrant multiplication, is believed by the author to be novel, and is suitable for the synthesizer application. The frequency of the multiplication is limited to the synthesizer application, but can be extended readily using suitable fast switching devices. The second multiplier circuit is an analog computer realization of the time division multiplier.

In the switching mode operational amplifiers have a low slew rate and a large saturation time which limits the switching frequency to 25 kHz in the case ot the μA741. However the principle of the multiplier can be refined as shown on the EAI 180 analog computer realization.

The circuit of Diagram 4.4.A. shows a voltage controlled mark to space oscillator. There is positive feedback through R_3 and R_4 from the output of the oscillator v_o, to the noninverting input of the integrated circuit v_+. The inverting input is connected to the same output voltage through a first order lag network, and also to the control signal input v_{i1} through R_1. The ratio of the time that the output voltage v_o is at the positive rail to the time that it is at the negative rail is determined by the control signal input. This relationship can be determined as follows.

The charge on the capacitor during one cycle is;

$$q_c(t) = \int_0^T i(t)dt$$

where i(t) is the current into the capacitor as shown on diagram 4.4.A. and T is the time of one cycle

The waveforms corresponding to the circuit of Diagram 4.4.A. are given in Diagram 4.4.B. Waveforms of $v_o(t)$, $v_i(t)$,$v_-(t)$ and $v_+(t)$ are given.

Separating the components of current into the capacitor;

$$q_c(t) = \int_0^T (v_o(t) - v_-(t))/R_2 \, dt + \int_0^T (v_{i1}(t) - v_-(t))/R_1 dt$$

where R_2 is the resistance between the capacitor and the output, and R_1 is the resistance between the capacitor and the control signal input

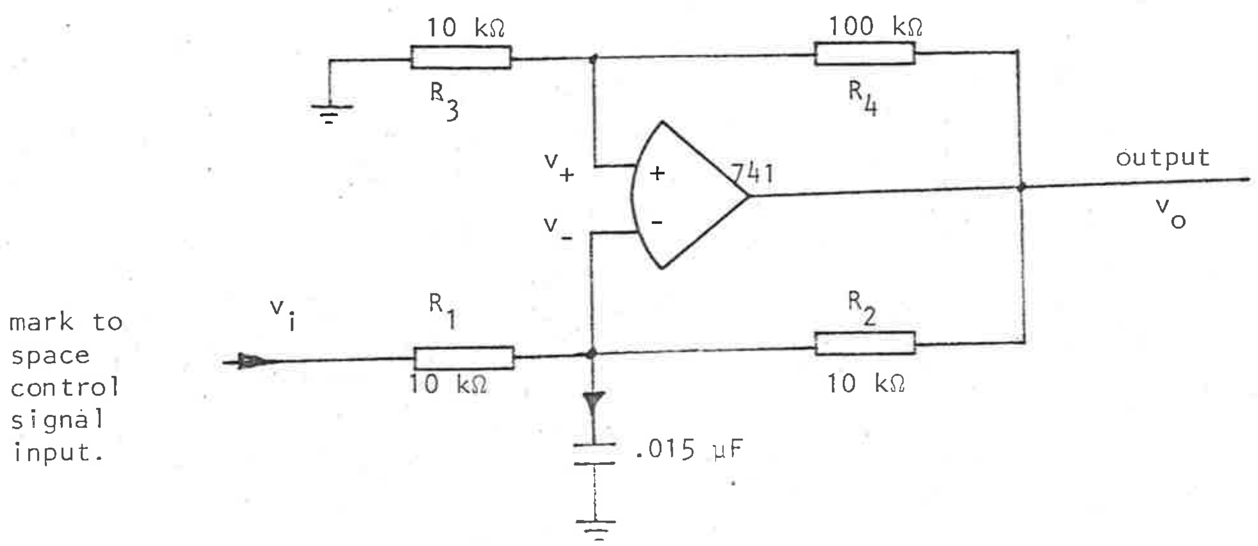


DIAGRAM 4.4.A. Voltage controlled mark to space oscillator.

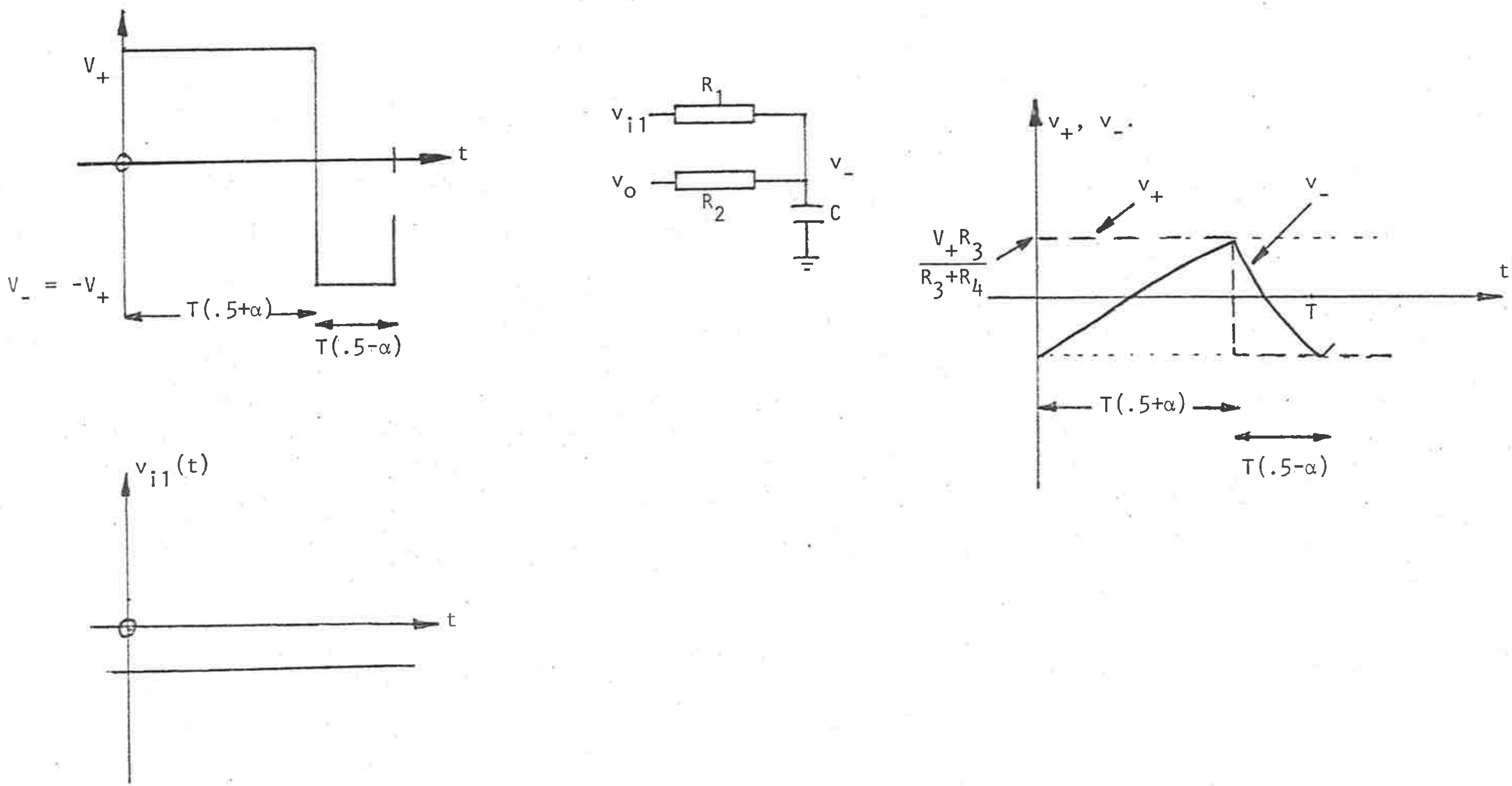


DIAGRAM 4.4.B. Voltage waveforms of the circuit of diagram 4.4.A.

Diagram 4.4.C. gives the simplified circuit for the switched analog inverter. When the analog gate in the inverter is short circuited, v_+ is at reference potential, and the circuit is a normal inverting amplifier. It has a gain of -1 in this case. The analog switch is a field effect transistor with a resistance during conduction of approximately 100Ω. This is .1% of the 100 kΩ resistor between the input voltage v_{i2} and the inverting input, so this resistance does not introduce a significant error. The common mode currents to the integrated circuit inputs are* 500 μA max for the μA741 and this gives an output voltage error under this condition which corresponds to this current through the two 4.7 kΩ resistors connected to the inverting input in parallel. The corresponding voltage error on the ouput is 1mV, and this can be avoided in precision applications of the multiplier by using integrated circuits with a lower common mode input current.

When the analog switch is switched off the resistance of the field effect transistor is now several orders of magnitude larger than the 100 kΩ resistor. Now the inverting input is at the same potential as the signal input because the current through the 100 kΩ resistor is very small. In this case the input currents to a 741 operational amplifier gives a 20 mV offset at the output. The large gain of the operational amplifier ensures that the noninverting input is at the same potential as the inverting input, which is the same as the input signal. No current flows through R_1, and hence no current flows through R_2. The stage now has a gain of +1.

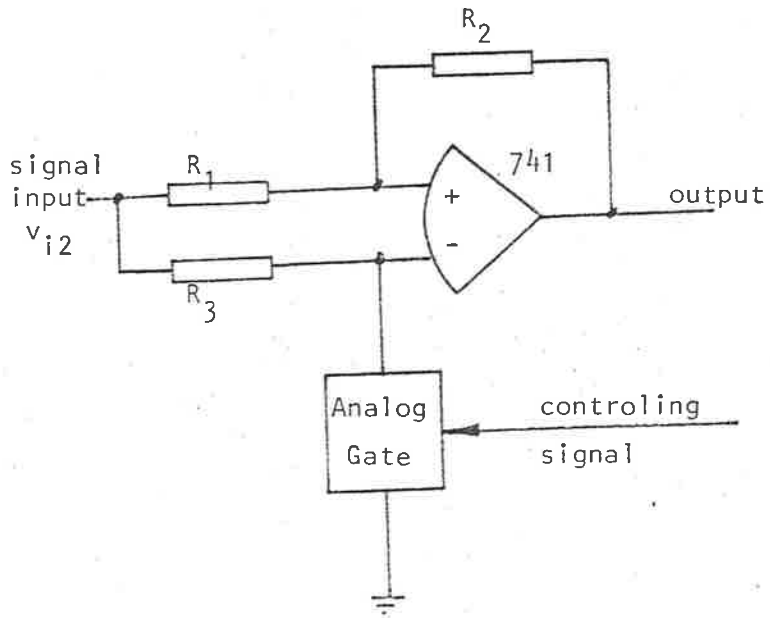


DIAGRAM 4.4.C. Simplified circuit for the switched analog inverter.

Hence $q_c(t) = \int_0^T v_o(t)/R_2 dt + \int_0^T v_{i1}(t)/R_1 dt + \int_0^T v_-(t)\{1/R_1+1/R_2\}dt$

The average value of $v_-(t)$, the voltage on the capacitor, over one cycle is very nearly zero, and also the nett charge into the capacitor over one cycle is zero, hence;

$$\int_0^T v_o(t)/R_2 dt = -\int_0^T v_{i1}(t)/R_1$$

The output of the operational amplifier switches between the positive rail voltage V_+ and the negative voltage rail $-V_+$.
Let the time the output is at the positive voltage rail during each cycle be $T(.5 + \alpha)$, and the time that the output is at the negative voltage rail be $T(.5 - \alpha)$, where T is the duration of each cycle.
It follows;

$$\int_0^{T(.5+\alpha)} V_+/R_2 dt + \int_0^{T(.5-\alpha)} -V_+/R_2 dt = \int_0^T v_{i1}(t)dt$$

Assuming that the input signal has frequency components which are low compared with the switching frequency, then $v_{i1}(t)$ is approximately constant during one cycle of the oscillator. Hence;

$$\frac{V_+}{R_2}(T(.5 + \alpha) - T(.5 - \alpha)) = v_{i1} . T/R_1$$

$$\alpha = v_{i1}\left[\frac{R_2}{2V_+R_1}\right]$$

This means that the average value of the output switching waveform is proportional to the value of the input signal v_{i1}. This waveform is now used to operate an analog inverter which has the second signal to be multiplied at its input.

Diagram 4.4.D. gives the overall block diagram of the multiplier, and Diagram 4.4.E. gives the circuit diagram.

The second signal input v_{i2} is noninverted for $T(.5 + \alpha)$ of each cycle, and is inverted for $T(.5 - \alpha)$ of each cycle of the mark to space oscillator. Since;

$$\alpha = v_{i1} \cdot \frac{R_2}{2V_+ R_1}$$

the average value of the voltage output of the inverter can be found in terms of the two input voltages.

$$\begin{aligned} v_{out} &= v_{i2}(t)(T(.5+\alpha)) - v_{i2}(t)(T(.5-\alpha)) \\ &= .5v_{i2}(t)T + \alpha v_{i2}(t)T - .5v_{i2}(t)T + \alpha v_{i2}(t)T \\ &= 2v_{i2}T\alpha \\ &= v_{i1}(t).v_{i2}(t)/V_+ \quad \text{when } R_2 = R_1 \end{aligned}$$

A low pass filter is required to filter the high frequency components of the inverter. Although the time delays in the linear operational amplifier limit the frequency response of this system they do not introduce errors within the passband. A time lag in the switching device does not alter the mark to space ratio which is determined by the nett charge on the capacitor over each cycle. However assymmetry in the supply voltages does alter the mark to space ratio and hence the output. If the negative rail is at a slightly lower voltage than the positive rail the time taken to supply the negative charge is longer than it would otherwise be, and therefore the mark to space ratio is changed. A circuit which uses an inverter to obtain these voltages in conjunction with voltage comparators is shown on Diagram 4.4.F. This was realized on an EAI 180 analog-digital computer, and in this case the slew rate of the integrator still limits the frequency response of the multiplier. The

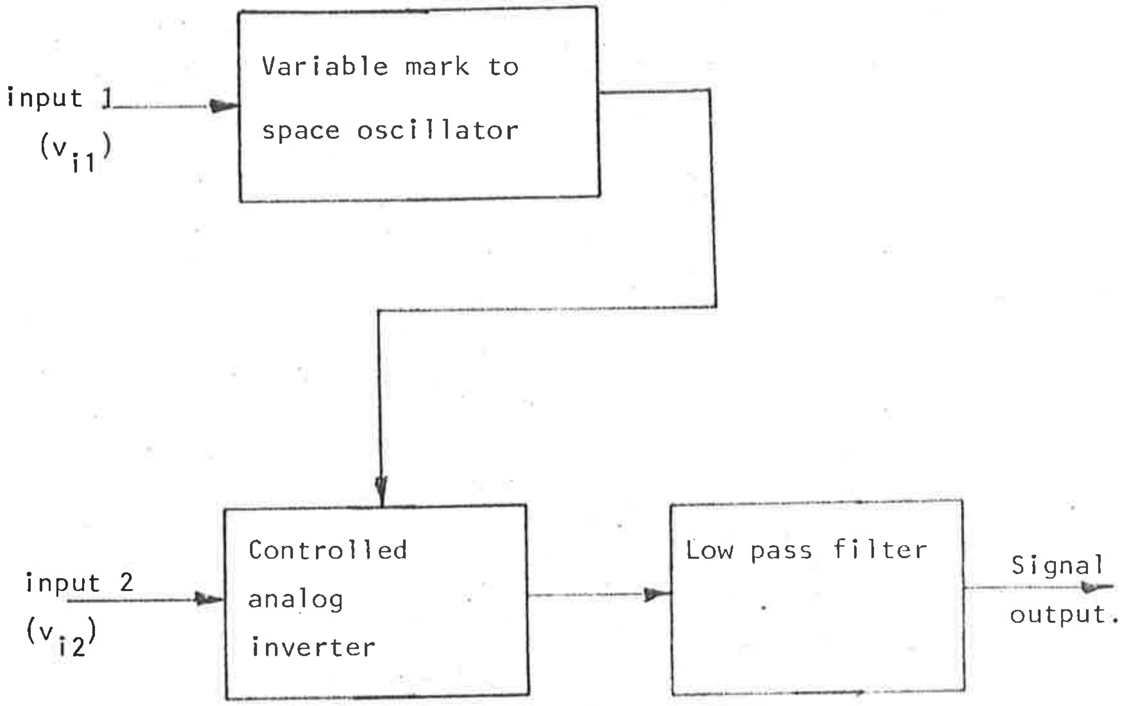


DIAGRAM 4.4.D. Block diagram of the analog multiplier

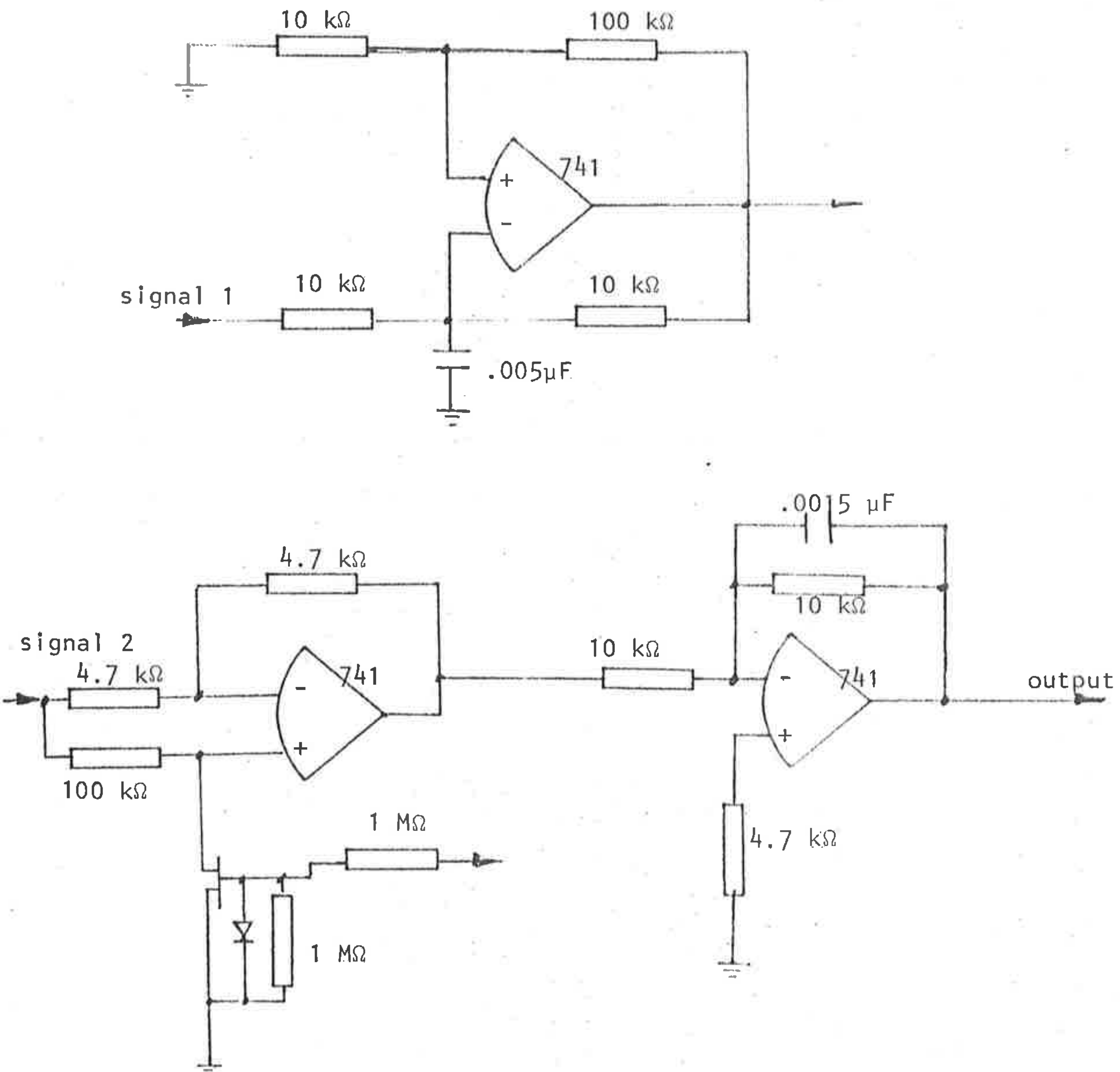
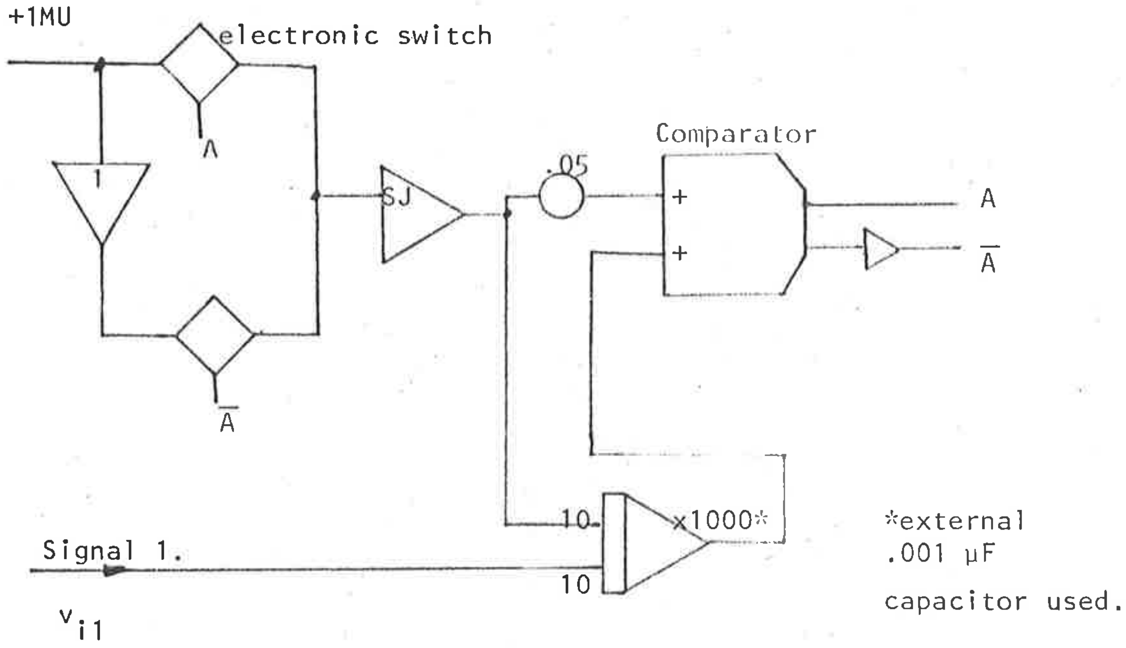
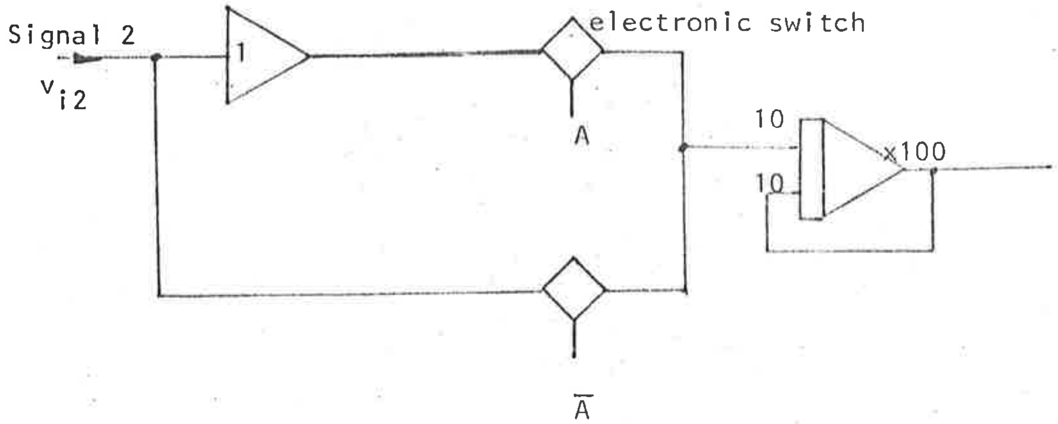


DIAGRAM 4.4.E. Circuit diagram of the multiplier



Oscillator



Inverter and Filter

DIAGRAM 4.4.F. EAI 180 Analog-logic computer realization of the time division multiplier.

original system uses only three integrated circuits and is quite adequate for the synthesizer application.

With this type of multiplier, once the variable mark to space oscillator is obtained any number of signals may be multiplied with very little additional hardware. This is useful if an application requires the multiplication of a number of signals by one given signal. The generation of harmonics from a sinusoid is one such application. The approach is also very valuable in application requiring higher linearity than can be provided by the present analog multiplier integrated circuits. Synchronous filters for spectrum analysers have this type of requirement.

4.5 Harmonic Generator

The harmonic generator is designed to add high frequency harmonic components to the input signal. This is achieved using the highly nonlinear transfer function of diagram 4.5.A. The harmonics added depend upon the waveform of the input signal and its amplitude. An arbitrarily small input signal will have a linear transfer function because of small signal linearization, and in this case no harmonics will be added. A squarewave input will also produce a squarewave output. Again the harmonic components will remain unchanged, though the relationship between the input and output signals' magnitudes will vary as the input signal's amplitude is changed.

A triangular input waveform does produce harmonic components in the output. The dominant harmonic depends upon the magnitude of the input signal as shown in diagram 4.5.B.

The transfer function of diagram 4.5.A was realized using a transistor to give the high gain required for the transitions in the transfer relationship. 20 transistors were required to give the 20 transitions as shown in the circuit diagram in diagram 4.5.C.

4.5.1 Operation of the Harmonic Generator Circuit

The input signal is buffered and sets up a series of 20 voltages in the voltage divider of series connected 100 Ω resistors. These 20 voltages range from + 6 V to - 6 V and are at equal intervals. As the potential of the midpoint of this resistive chain is altered by the input signal, the transistors are switched on or off. The collectors of these transistors control the current to the differential operational amplifier. When the currents on the inverting input

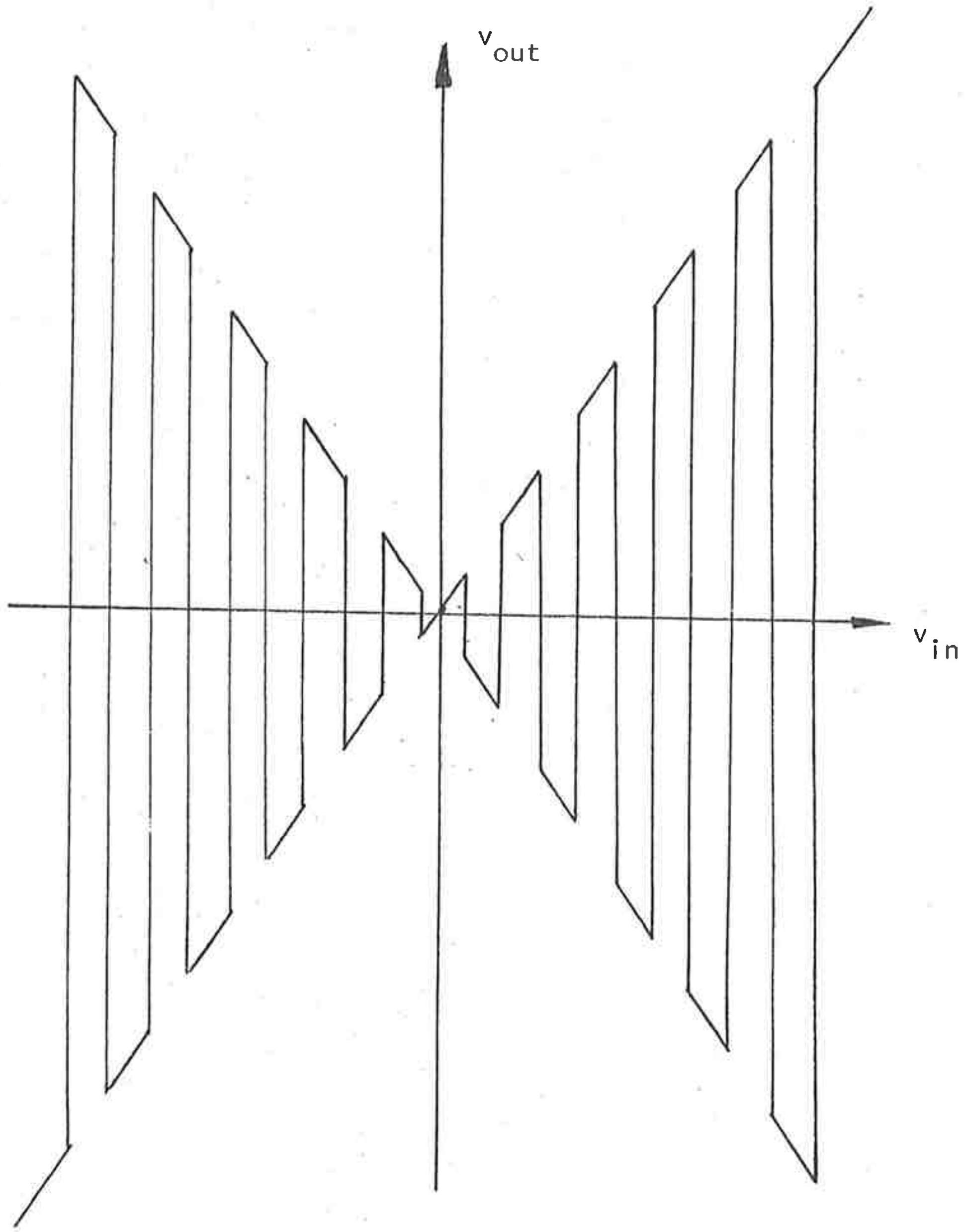
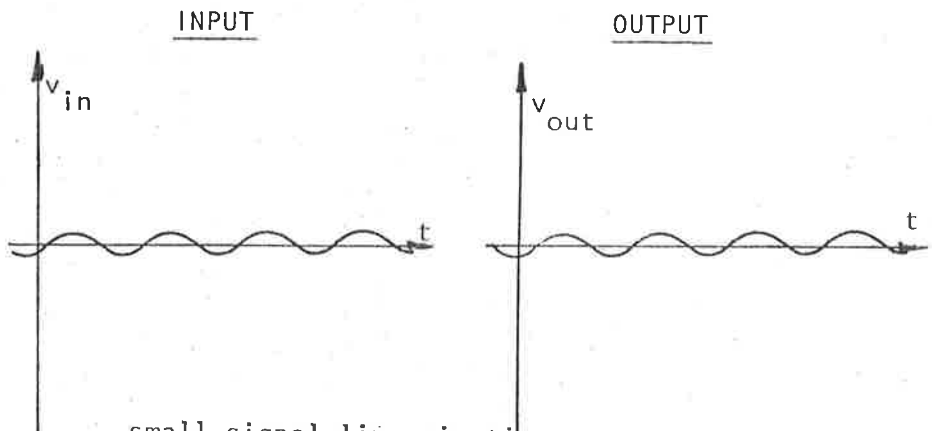
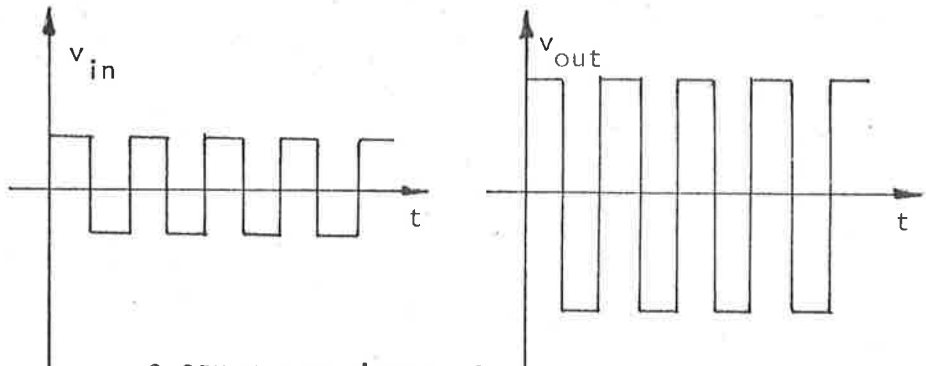


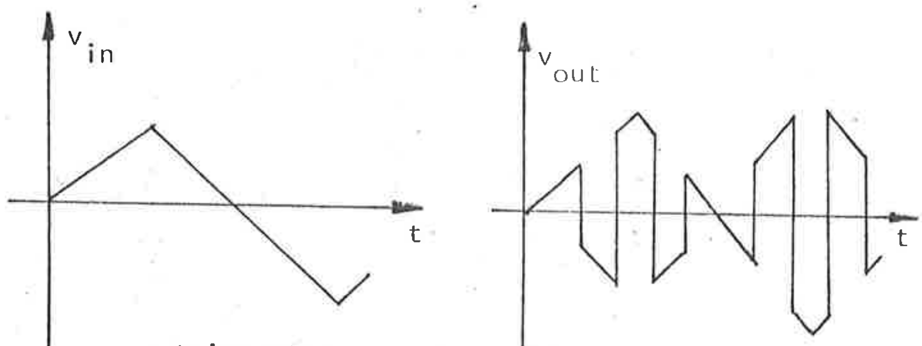
DIAGRAM 4.5.A. The nonlinear transfer function of the harmonic generator.



small signal linearization



a squarewave input gives a squarewave output



a triangular wave input gives harmonics in the output. The larger the input amplitude the higher the frequency of the harmonic components.

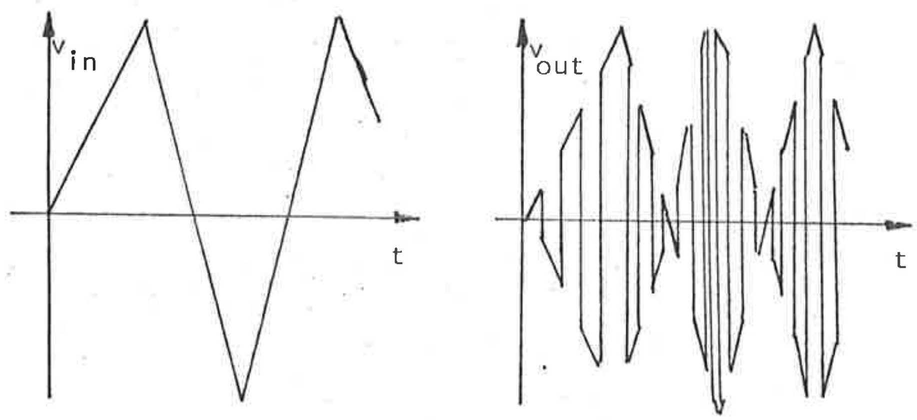


DIAGRAM 4.5.B. Some input to output relationships for the harmonic generator.

equals the current on the noninverting input then the output is very nearly zero. In general there is a slight offset potential of these inputs, but this is made small by the 1 kΩ resistors shunting the inputs to ground. This means the output of the differential amplifier gives the required input to output relationship.

The value of the resistor combination in the collector for each segment can be calculated knowing the currents required at the input of the differential operational amplifier.

Input Voltage	Required Output Voltage	Nett Input Current μA	Additional input Current: inverting input μA	Additional input Current: noninverting input μA	Total Resistance:* inverting	Total Resistance:* non-inverting
.3	.3	.7		.7		2.1 MΩ
.9	-.9	-1.9	2.6		580 kΩ	
1.2	1.2	3.2		5.1		290 kΩ
1.5	-1.5	-4.5	7.7		190 kΩ	
2.1	2.1	5.7		10.2		150 kΩ
2.7	-2.7	-7.0	12.7		118 kΩ	
3.3	3.3	8.3		15.3		98 kΩ
3.9	-3.9	-9.5	17.8		84 kΩ	
4.5	4.5	10.9		20.4		74 kΩ
5.1	-5.1	-121	23.0		65 kΩ	

*15 V is applied to this total resistance when the transistor is switched off.

The total resistance determines the current to the inverting or the noninverting input of the differential amplifier. This includes the collector resistor of the transistor as well as the resistor between the collector and the inverting or noninverting inputs. The collector resistor and the series resistor to the virtual earth can be designed as follows.

Total Resistance Required kΩ		Collector Resistor kΩ	Series resistor to virtual earth. Inverting kΩ	Noninverting kΩ
	2100	100		2000
580		100	470	
	290	100		180
190		100	89	
	150	47		100
118		47	68	
	98	47		47
84		47	39	
	74	33		39
65		33	33	

The circuit is now given in diagram 4.2.C.

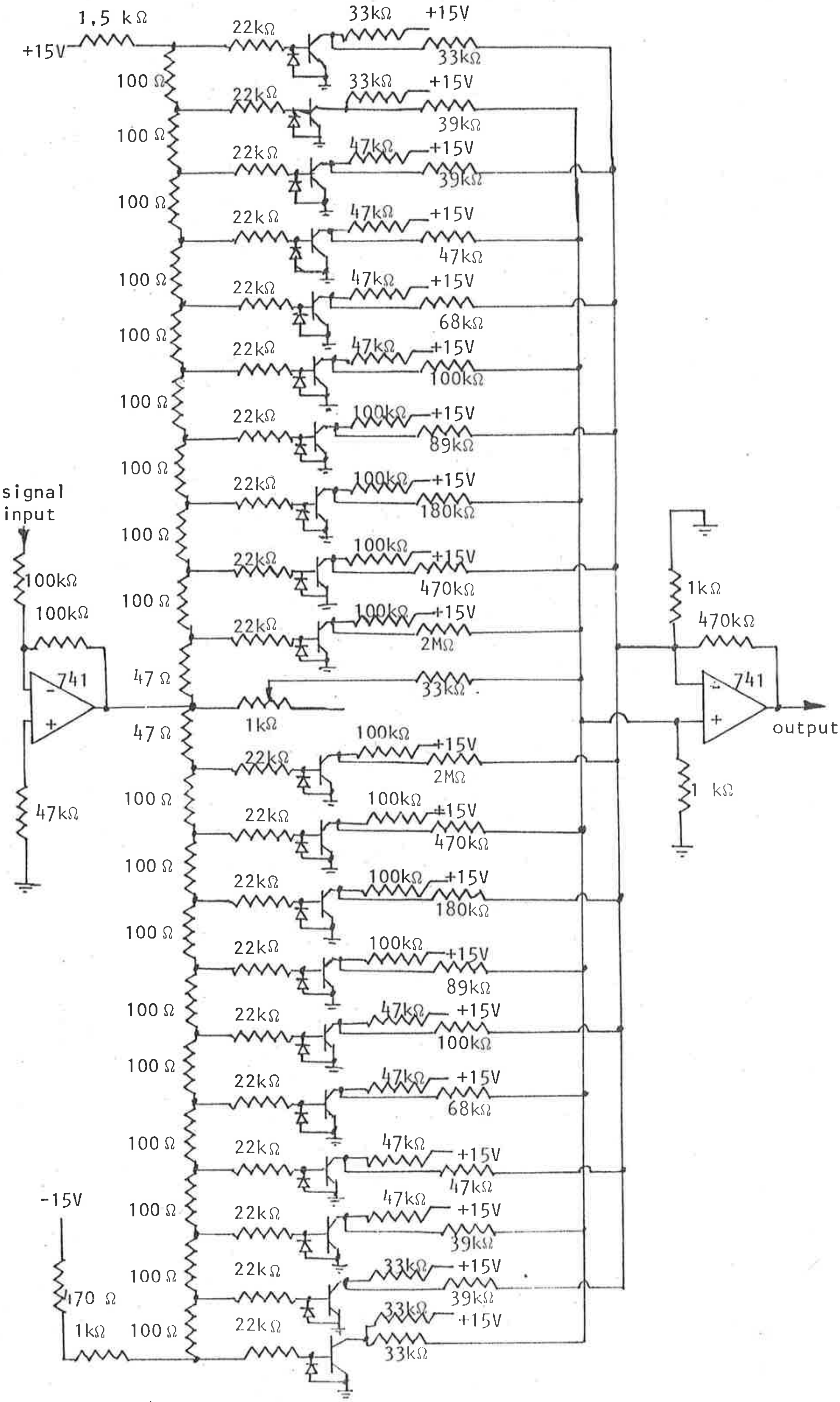


DIAGRAM 4.5.C. Circuit of the harmonic generator.

Two refinements to the Harmonic Generator became apparent after it was built. When the input to output transfer function $v_2(v_1)$ is odd (where v_2 is the output voltage and v_1 is the input voltage), then the number of transistors may be halved using the circuit of diagram 4.5.D. The even simpler arrangement of diagram 4.5.C produces the even transfer function where $v_2(v_1) = v_2(-v_1)$, which is quite satisfactory for this application.

Although the transfer function is grossly nonlinear, it was found that the output waveform still had considerable components of the fundamental frequency of the input signal. This is because for input waveforms other than the ramp or triangular waveform the transitions are not at equal time intervals. An adjustment for cancelling the input signal was incorporated to emphasize the harmonic content. Unfortunately this adjustment depends upon the input signal, however it was still found to be useful.

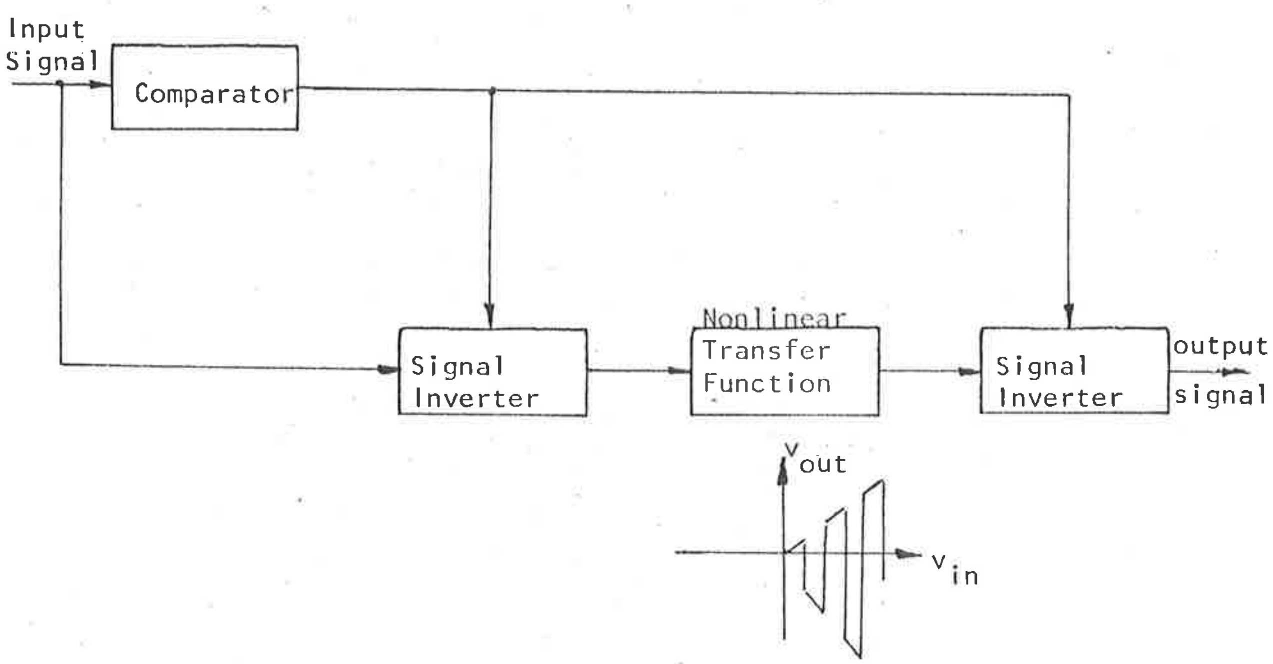


DIAGRAM 4.5.D. Block Diagram for realizing an odd input to output transfer function with a saving in the number of switching elements.

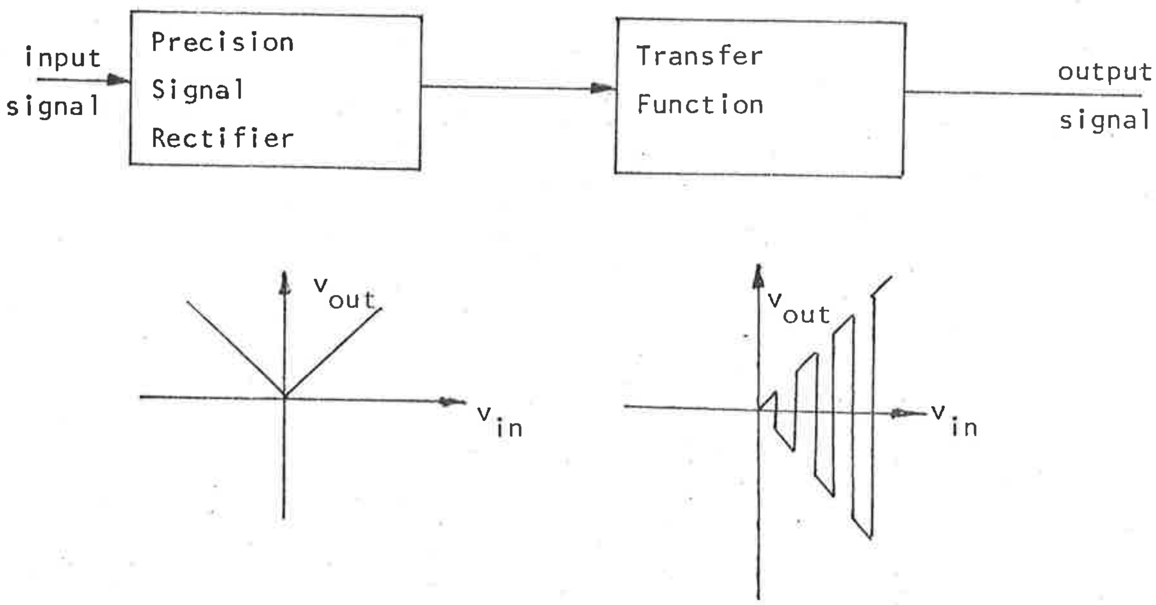


DIAGRAM 4.5.E. Block diagram for realizing an even trensfer function, again with fewer switching elements.

4.6 Reverberation Unit

An echo occurs when a sound is heard after some time delay. Reverberation occurs when a number of echos are present but no one echo need be predominant. Usually sounds are transmitted and received in the presence of objects which give reverberation effects. In fact, hearing sounds without reverberation such as within an anechoic chamber is quite an unusual experience. Therefore the reverberation effect is quite important in a music synthesizer.

Reverberation and echos both involve the storing of sound information. The unit investigated for the synthesizer was an acoustic delay line which used a loudspeaker and a microphone placed at the ends of a length of pipe. The transducers contributed an unacceptable amount of distortion, and the pipe severely attenuated the frequency components above 1 Kz. It was decided to use a commercially available mechanical spring delay line.

Electrical means of storing sufficient information have recently become readily available and at an acceptable price. The number of bits which would be required can be calculated.

Number of bits per analog sample (1%) = 7

Theoretical minimum sampling rate (Shannon)

for a cutoff frequency of 10 kHz = 20 kHz

Practical sampling rate = 30 kHz

Duration of the storage of the information = .2 s max

Number of bits required = $7 \times 30 \times .2 \times 10^3$ = 42 k bits

This amount of digital storage is now available but was prohibitively expensive when the design was considered. Modern charge control devices will be very suitable for this application in the near future. However at the time the design was finalized it was decided to use a commercially available mechanical spring delay line.

4.7 Position Control

Composing techniques often require a spatial effect. Groups of sounds can be moved by altering their relative amplitudes in two or more channels. It would be a considerable advantage if this process could be controlled by a voltage so that the apparent movement of the sound could be related to the output of any control unit, or even a signal.

Consideration was given to presenting the musician with a most flexible system of voltage controlled attenuators. The front panel was designed so the position control attenuator system was convenient to operate and yet give flexibility.

Field effect transistors can be used as inexpensive signal attenuators. If the gates of a P channel field effect transistor and an N channel field effect transistor are connected to the same control voltage, a given control voltage increment increases the attenuation in one channel and decreases it in the other. However the relationship between an increment in the gate voltage of a field effect transistor and their small signal resistances vary widely. It is possible to choose field effect transistors from a large number so that they have similar characteristics. However the use of selected components makes the maintainence of the device very difficult. An alternative approach is to provide adjustments to give a linear transposition of the control voltage:

$$v_{gate} = k_1 v_{control} + k_2 .$$

where k_1 and k_2 are constants which depend upon the adjustments of the trim potentiometers.

The source to drain resistance of the field effect transistors depends upon the voltage between the source and the drain as well as the gate voltage. This nonlinearity means that the voltage across the field effect transistor should be minimized so as to minimize the distortion products. The extremely large signal power gain of linear operational amplifiers can be used to make this signal voltage small. A voltage gain of 100 is used so that the signal voltage across the field effect transistor is 50 mV or less. Even larger gains could have been used with A.C. coupling. However with D.C. coupling the input current and voltage offsets of the operational amplifier become important. D.C. coupling has the advantage that there is no low frequency limitation of the attenuator system.

The field effect transistor can be used to shunt the signal, or in series with the signal to form an open circuit. The series connection is more suitable for this application. The resistance of the field effect transistor is extremely large (in excess of 10^9 Ω) whereas the resistance when conducting is typically 100 Ω. It is important that there is very little feedthrough the attenuator when it is blocking the signal; even with 60 db of attenuation detectable signal components are still present, and this significantly detracts from the usefulness of the attenuator. The off resistance of the field effect transistor is so large that the required attenuation can be achieved if the series connection is used.

The position controller has four signal inputs and two signal outputs. Two attenuator pairs are connected to two control voltages. This allows independent control of two sounds. The position of these sounds could be determined by touching a keyboard, or by using a lever. Each of the four attenuators can be used independently. In this case, as one sound is reduced in one channel, another sound increases in the other. There is a 4 x 4 signal switching matrix provided with the position controller to give this flexibility.

The circuit of the position controller is given in diagram 4.7.A

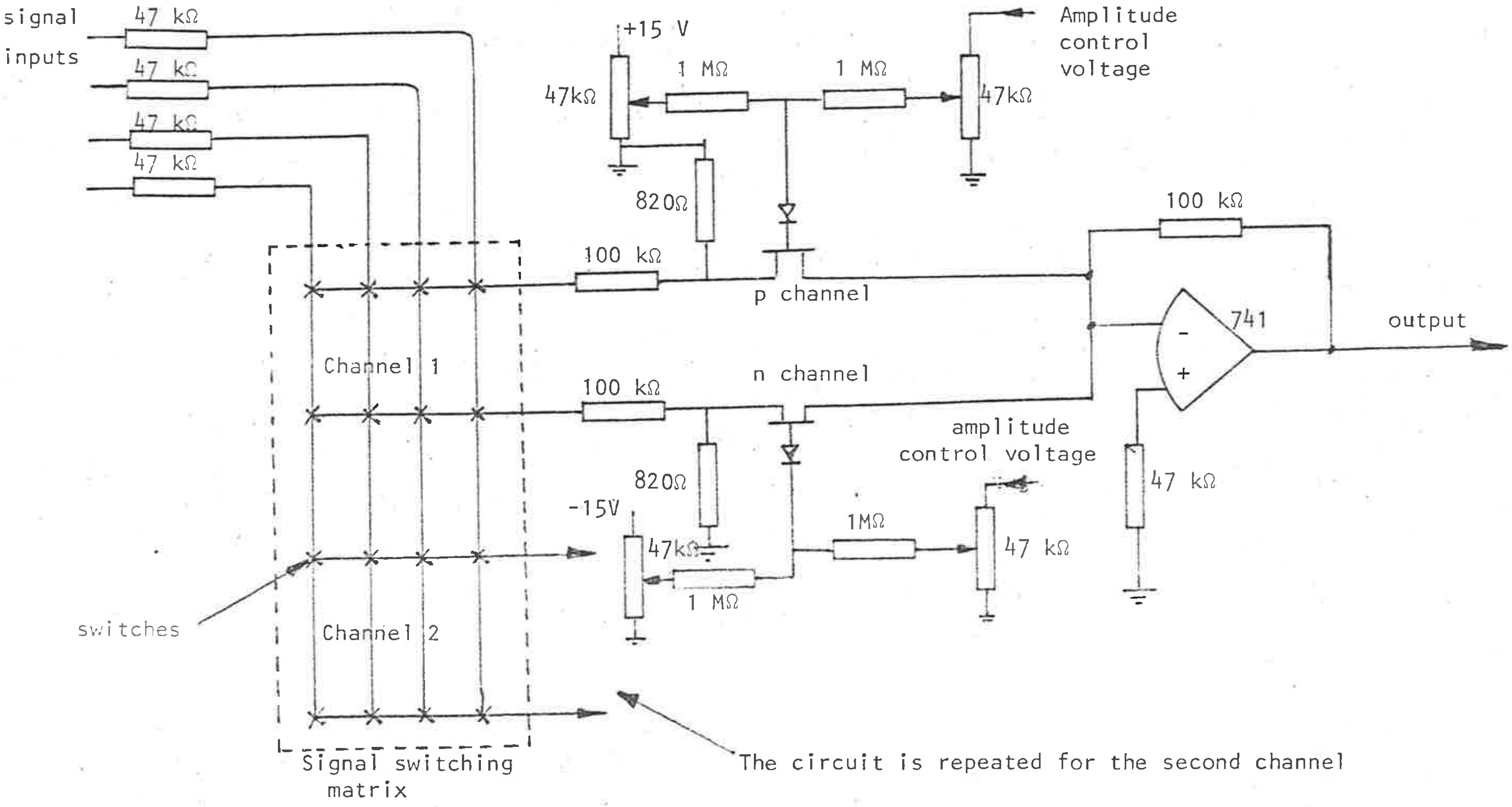


DIAGRAM 4.7.A. Circuit Diagram of the Position Controller.

Pan control.

4.8. Frequency Divider

4.8.1 Frequency components produced by nonlinear circuits.

Frequency components introduced by amplitude distortion tend to be higher than those in the input signal. The object of the Frequency Divider sound manipulator is to provide frequency components which are submultiples of the input frequencies. The Harmonic Generator (4.5.) unit uses a grossly nonlinear transfer function to produce predominantly high frequency components. If the input signal to a nonlinear system is periodic, and there is a conservative nonlinear relationship between the input voltage, then the frequency components introduced by the nonlinear element must be multiples of the fundamental input frequency. This is because all the sinusoidal components in the output signal must be periodic in the same peroid as the input signal. In some special cases the harmonic components present at the input can actually be reduced. This is the case when a triangular waveform is applied to a sinewave shaper circuit. As a general rule, however, nonlinear circuits increase the harmonic content.

When a number of frequency components are present at the input of a nonlinear element intermodulation of the frequency components occurs. This produces the sum and the difference frequencies of the input sinusoidal components, and the output can contain frequency components which are lower than any input frequency. However these frequency components predominate only with special combinations of input signals and nonlinear functions. The objective of the Frequency Divider sound manipulator is to provide considerable components of submultiple frequencies for almost all input waveforms. It follows that a conservative relationship between the input voltage and the output voltage

contd/..

does not suit the design objectives of the Frequency Divider unit.

4.8.2 Use of Memory with the Frequency Divider.

The submultiple frequency components can be produced using a memory element such as a flip-flop. Digital division provides large components of submultiple frequencies. However the output of these division circuits is a square wave of constant amplitude. For a musical synthesizer this sound would rapidly become very monotonous; the harmonic components would never change and the amplitude would also be fixed. The design requirements of the sound manipulator dictate that many sounds should be available, and that these sounds should relate to the input sound in a useable way. This would allow the musician to use the sound manipulator creatively. With the Frequency Divider unit these requirements are realized by using the digital signal, which is a submultiple of the input frequency, to invert the original input signal. If the digital signal is at logic '1' the input signal remains unchanged at the output, but is the digital signal is at logic '0' then the input signal is inverted. This means the amplitude of the input signal determines the amplitude of the output, and that input signals with fewer harmonic components give outputs with fewer harmonic components, though some harmonic components are always added.

4.8.3 Effect of the Frequency Divider in the time domain.

The Frequency Divider unit has independent provision for division by 2, 3, and by 4. The corresponding output signals can be combined independently. For a sinusoidal input the time waveforms of Diagram 4.8.A. are obtained.

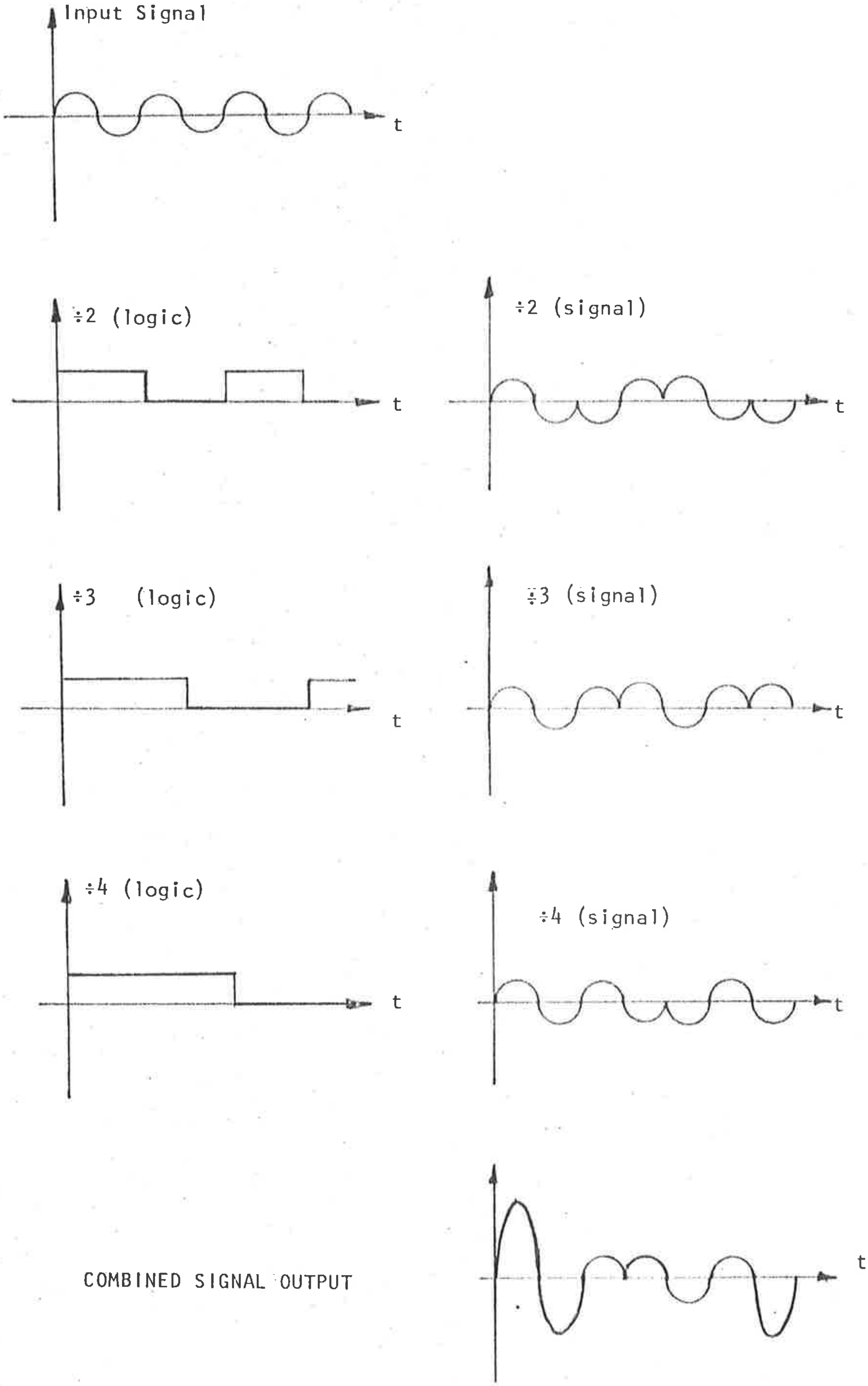


DIAGRAM 4.8.A. Logic and signal waveforms of the Frequency Divider.

4.8.4 Frequency Domain Representation for periodic signals.

Since $$2\, R_e|Z_1|R_e|Z_2| = R_e|Z_1Z_2 + Z_1Z_2^*|$$

it follows that for two periodic signals,

$$A(t) = \mathrm{Re}\left|\sum_n a_n e^{jw_nt}\right| \quad \text{and } B(t) = \mathrm{Re}\left|\sum_m b_m e^{jw_mt}\right|$$

the product,

$$2\,A(t).B(t) = \mathrm{Re}\left|\sum_n a_n \sum_m b_m e^{j(w_n+w_m)t} + \sum_n a_n \sum_m b_m e^{j(w_n-w_m)t}\right|$$

If A(t) is the periodic input signal to the Frequency Divider unit and B(t) is the squarewave of submultiple frequency of the fundamental component of A(t), then the resulting frequency spectrum can be found as shown in Diagram 4.8.B. Subharmonic components of A(t) are present with frequencies and magnitudes which depend upon the magnitude of the harmonic components in the input signal, A(t). This frequency domain representation of the signal multiplication can be used to show that the characteristics of the input signal is preserved by the Frequency Divider unit.

The block diagram and the circuit diagram of the Frequency Divider unit are given in Diagrams 4.8.C. and 4.8.D. respectively.

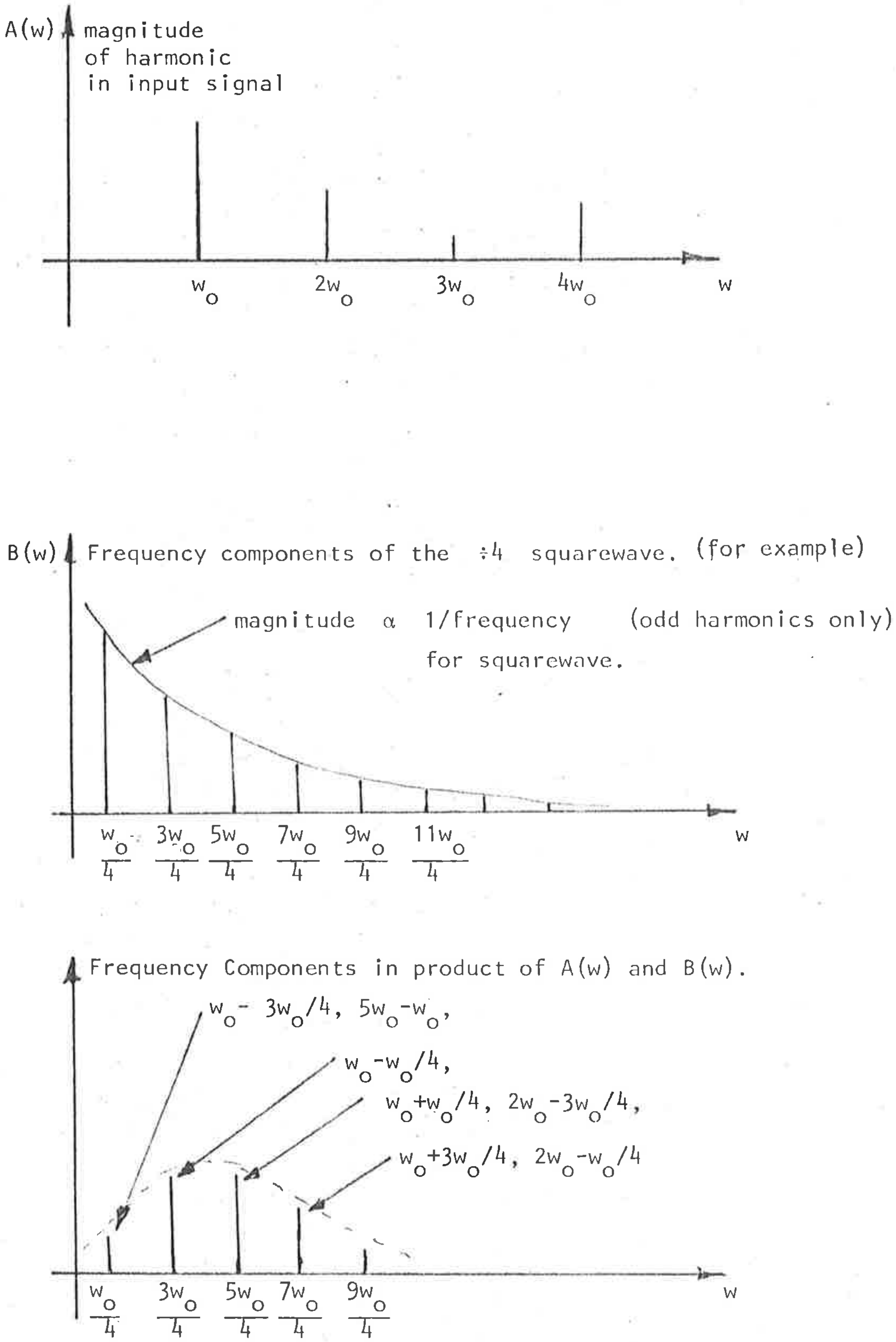


DIAGRAM 4.8.B Diagram of the frequency spectra of the input signal the generated submultiple, and the corresponding product.

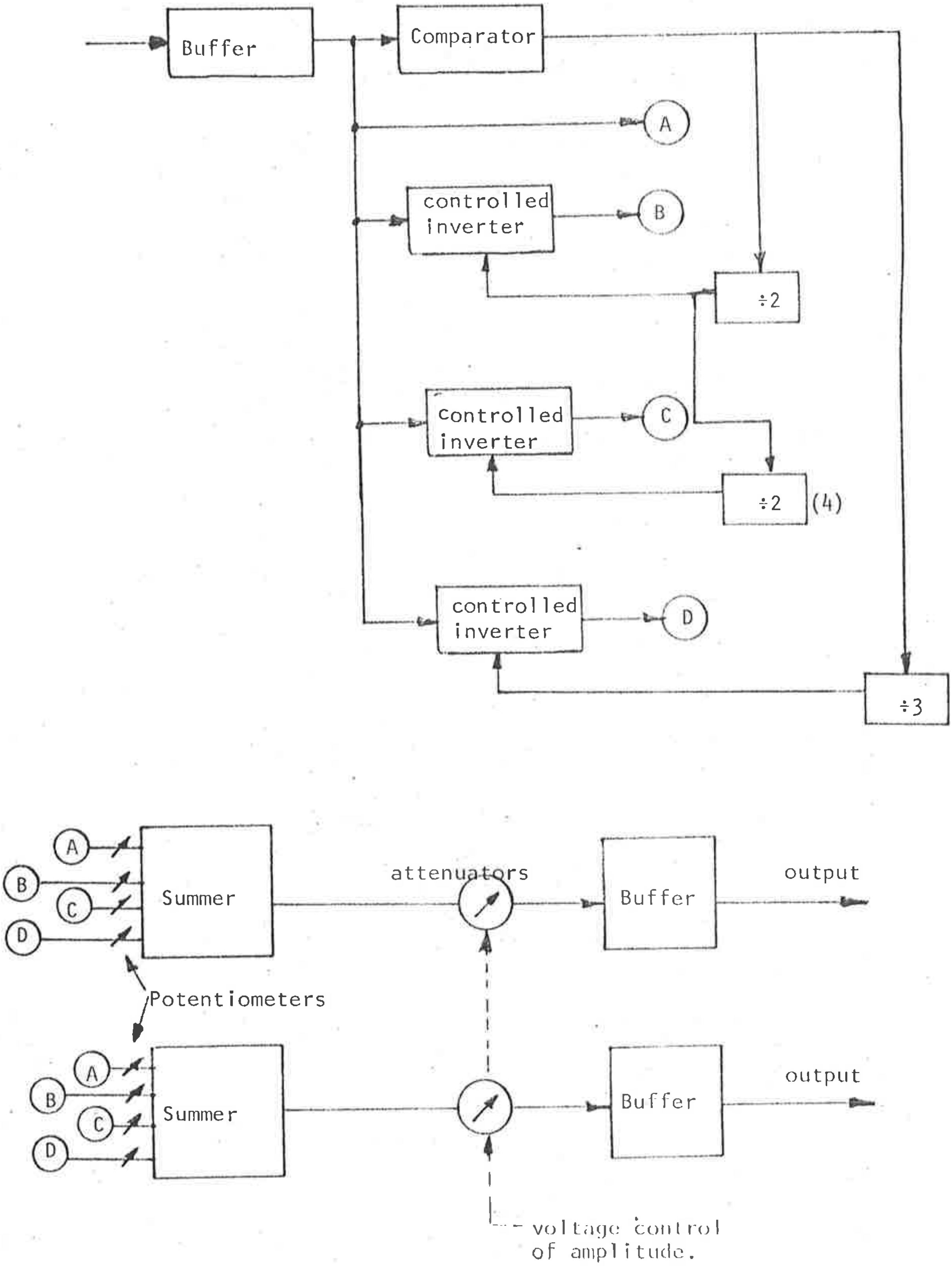
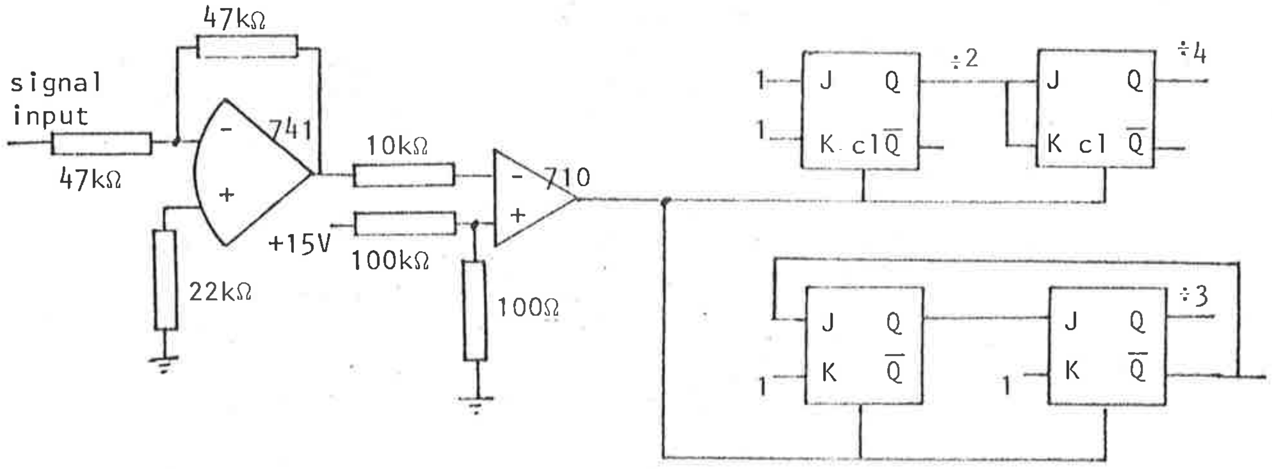
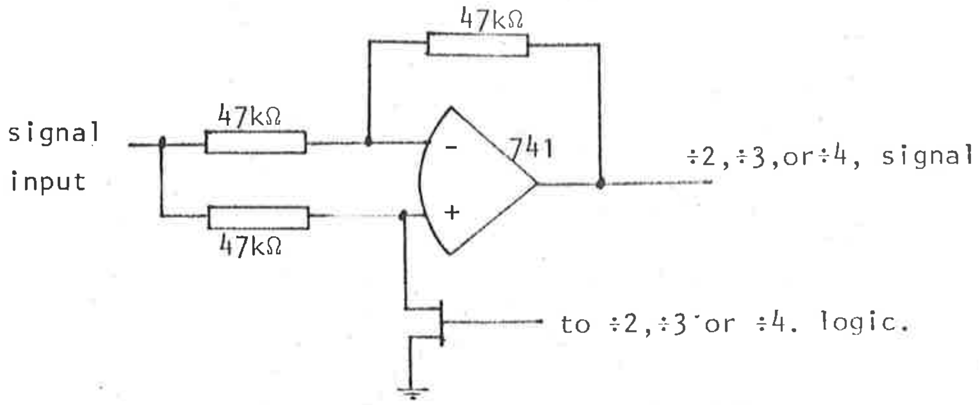


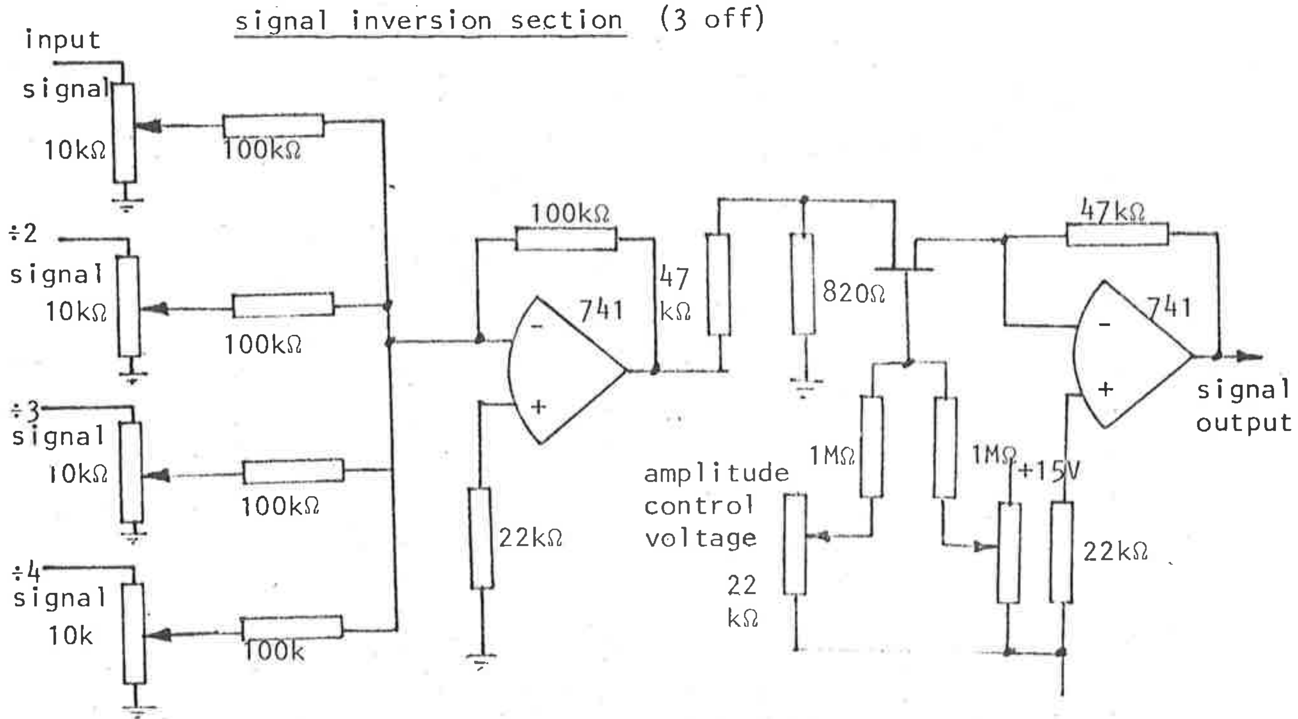
DIAGRAM 4.8.C. Block diagram of the Frequency Division sound manipulator



logic section



signal inversion section (3 off)



summer and voltage controlled amplitude section (2 off)

DIAGRAM 4.8.D. Circuit diagram of the Frequency Divider Sound Manipulator

4.8.7 Notes on the circuit design.

It was found that a 15 mV offset on the comparators was necessary to stop high frequency triggering of the comparator when no signal was present at its input (except for very low level noise components). Otherwise this rapid switching of the comparator output fed through to the analog section to some noticeable effect.

The two voltage summers are each fed by components of the four signals (input,÷2,÷3, and ÷4), and the amplitude of each component can be adjusted independently. The field effect transistor attenuator is arranged so that as one amplitude level increases the other falls. This is achieved using an N channel FET in one output, and a P channel FET in the other. This means that if the outputs are added on the signal switching matrix (Ch 2.), the control voltage can be used to change from one combination of the outputs to the other.

4.8.8 Operation of the Frequency Divider.

With each of the 2, 3, and the 4 sections in use the output's fundamental frequency is 1/12th of the fundamental frequency of the input. This component gives a very distinct acoustic effect.

All the sound manipulators can be used to produce a control voltage. They could be used in conjunction with a signal source and a control unit. The effect of the frequency divider in a control voltage differs considerably from otherwise available units, and again this difference is reflected in the corresponding sound which is produced.

NOTES FOR CHAPTER 4

1. See references 24 and 25 for the presentation of the Biquad Filter principle.

2. The technique of analysis used for the filter is from control system theory. This is given in references 14 and 26.

3. This multiplier principle is outlined for termionic valves in reference 26 p 5.16.

CHAPTER 5 CONTROL UNITS

5.1 Introduction

The function of a control unit is to relate a physical operation to a control voltage which in turn determines the sound output. The synthesizer is designed so the operator has access to a large number of sounds and, further, the operator is monitoring the sound as he is selecting it. Therefore the physical operations must occur at the same time as the sound itself. This means the control voltages must conveniently relate to the physical operations so that in turn the sound parameters can be conveniently related to the physical operations. It also follows that the controllers should use as many aspects of the operator's movement as practicable. An example of this design approach occurs with the Lever Controller. Instead of having one rotating potentiometer and therefore only one control voltage output, three can be arranged orthoginally to give three independent control voltages. Another example occurs with the Binary Keyboard. This keyboard is sensitive to pressure as well as to which combination of notes is depressed. The advantage of the corresponding ability to control amplitude using pressure is illustrated by the piano-forte.

The relationship between the physical movement of the operator and the corresponding control voltage is chosen so that a given change in position gives a given change in the control voltage. This minimizes the effects of noise and offset errors. The sound parameters are then related to the control voltage in a way which suits the ear. Frequency and amplitudes are logarithmically related. The output impedance of the voltage from each control unit is low, and these are combined using the control voltage switching matrix so that the total of the input voltages is obtained. This ensures a given control voltage is not changed when an additional control voltage is added, so that the control unit need not be readjusted as additional units are switched in. It also means care must be taken to avoid overloads of the control voltages.

The range of the control voltage output can be adjusted. It follows the sensitivity of the voltage variation to the physical movement can be chosen also.

Established keyboard instruments duplicate sound sources for each note, thereby allowing simultaneous operation of the keys. Although simultaneous operation of the keys of an electronic organ is provided, it is not avialable with this type of music synthesizer. With music synthesizers one sound source or group of sound sources is controlled by one control voltage, and a single control voltage cannot provide independent and simultaneous control of more than one sound source. However combinations of the keys of a keyboard can still be used to give more flexible control. An unnecessary restriction applies to the keyboard of the Moog synthesizer; only the lowest key depressed affects the control voltage. This means that simultaneous control by the other fingers is not used. With the Conventional Keyboard unit, a more convenient electrical circuit which uses only one switch per note gives a voltage average when more than one key is depressed. This means that semitones and other small frequency increments are available without additional complexity.

A more flexible Binary Keyboard was designed to give voltage sums as well as averages. When the control voltage is logarithmically related to an oscillator frequency, adding a given voltage increment corresponds to increasing the frequency by a constant interval or frequency ratio. Similarly, averaging the control voltage gives the geometric mean of the corresponding frequencies. The Binary Keyboard provides both averaging and adding of the control voltages when more than one key is depressed. This means $1027 = 2^{10} - 1$ control voltage levels are obtained using only 10 keys. Also, being physically smaller than the conventional keyboard, it is more convenient to obtain a control voltage from the applied pressure.

A Lever Controller was used to obtain analog control and hence sounds differing in character from the discrete levels of the keyboard controllers. The lever has three simultaneous outputs, and this is an extension of the slider controller of the Moog synthesizer.

With the Resistance Board Controller a board of resistance paper was made to give voltage outputs which varied with position. It is possible to choose the shape of the resistance paper so that a variety of potential gradients are available. Also, signal voltage can be superimposed at various places so that the corresponding sound outputs have different characteristics.

A Foot Controller was designed to begin to use other limbs. The feet of the operator are used to considerable advantage with some established musical instruments such as the organ, piano-forte and drums.

The control unit should obtain control voltages from any convenient source. This is illustrated by the Sound to Voltage unit. This has output voltages which are functions of the freuqnecy and amplitude of an input signal. Ths input signal could then be from any source.

If the input signal is from a microphone the training of the voice could be used to control the parameters of the sound produced by the synthesizer. Once the control voltage is obtained the relationships between the input signal's amplitude and frequency to the output signal's characteristics are very flexible.

A Control Voltage Manipulator was made to provide the differential and the integral of the control voltage. These are combined in adjustable amounts with the origional control voltage.

These seven units are now discussed in detail. The averaging and the summation of the control voltage on the Binary Keyboard, the lever controller with three degrees of freedom, the sound to voltage unit and the signal injection of the resistance board controller and the control voltage manipulator are believed to be novel.

5.2 Conventional Keyboard

The conventional keyboard of a piano is a very widely used method of selecting musical sounds. It has many notes, and in most instruments which use this keyboard more than one note can be played at one time. This means the mechanism for producing the sound must be duplicated many times, and this does not suit the synthesizer application.

The sound sources of a synthesizer do not produce more than one note at the same time and need only one control voltage to select frequency. The conventional keyboard used has 61 keys and this means any method of sensing the applied pressure would have to be duplicated many times. Therefore only one control voltage is provided by this control unit.

The circuit of the conventional keyboard is given in Diagram 5.2.A. The circuit is arranged so that there is a constant voltage interval between successive keys. The voltage levels of both of the extreme keys are both independently adjustable. Combinations of keys are used; if more than one key is depressed then the average voltage is obtained. This means intermediate frequencies such as semitones are available to the musician. The circuit requires only one switch per key, and a portomento control is provided. The portomento control corresponds to a first order lag in the control voltage, and the steady state control voltage is approached with a significant timeconstant.

The method of adjustment of the voltage levels corresponding to each of the keys illustrates an interesting aspect of control unit design. Th obtain independent control of each key it is necessary to provide an indenendent means of adjustment. For instance the increments on the keyboard of the Moog synthesizer can be adjusted independently, but to do this it is necessary to provide a potentiometer for each key. This large number of potentiometers is both expensive and difficult to set up. Even so, in the case of the Moog synthesizer the voltage levels which correspond to successive keys must be in monatonic order. To enable independent control of the voltage from each key it would be necessary to have higher quality potentiometers so they could be adjusted more accurately. If there are few controls for setting up the control unit then the unit cannot be used flexibly.

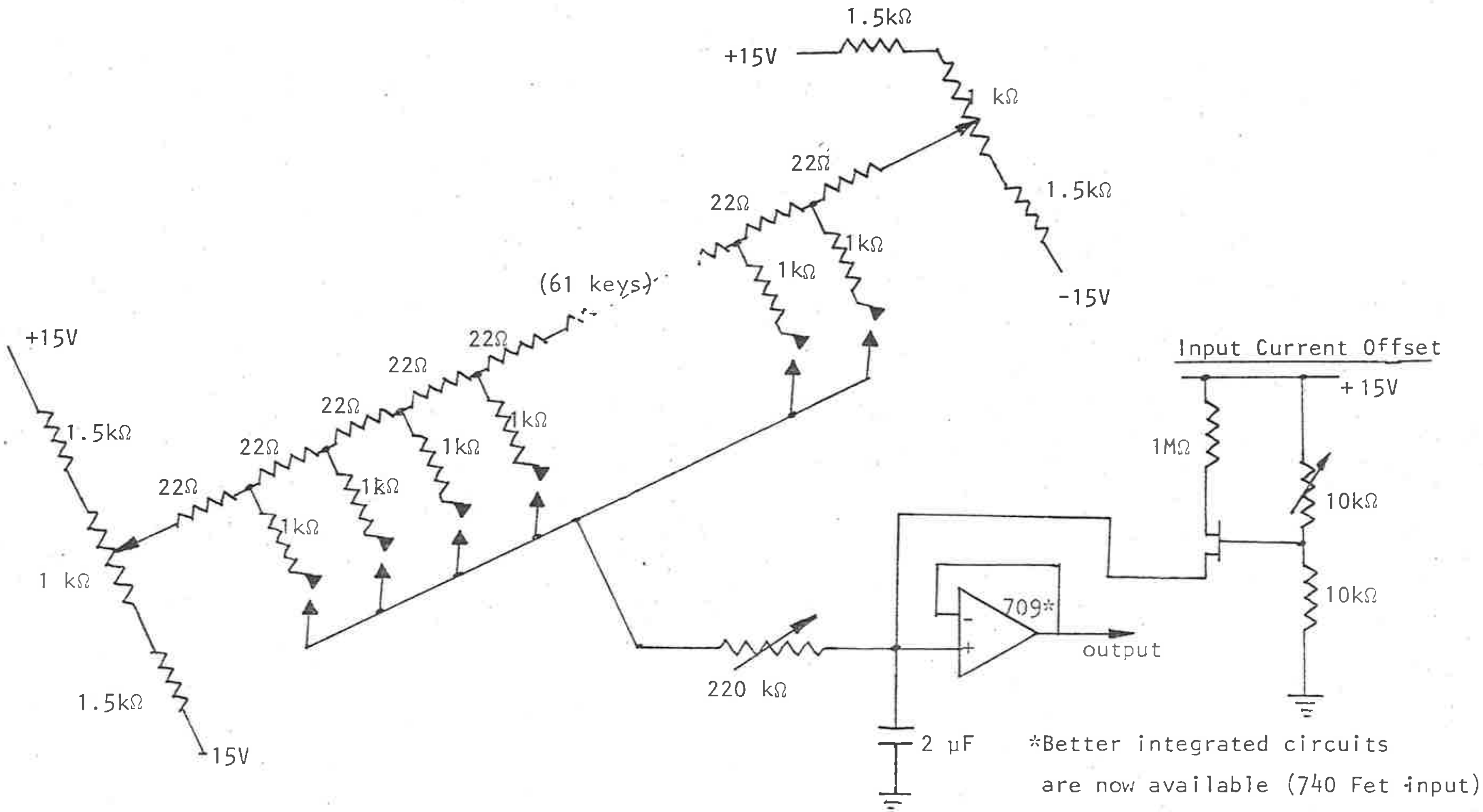


DIAGRAM 5.2.A. Circuit of the Conventional keyboard.

5.3 Binary Keyboard

The Binary Keyboard is designed specifically for the music synthesizer. It has only 10 keys which is a much smaller number than the 81 keys of a conventional keyboard. However the conventional keyboard was designed for instruments which allowed notes to be sounded simultaneously, a characteristic which is not easily obtained with electronic instruments. An electronic organ, which has a very different design basis from a synthesizer, requires the provision of individual notes for each of the keys. This represents a large investment. The use of voltage control in a sound synthesizer makes the simultaneous provision of notes even more difficult. Each voltage controlled oscillator has only one independent output.

Many conventional musical instruments can only provide one note at a time. This is true of a trumpet. The trumpet is an example of an established musical instrument where a few keys are used to choose between many frequencies. This is possible because the keys which determine the frequency of the trumpet operate so that combinations are important. The Binary Keyboard was designed to obtain this characteristic in a control unit for a synthesizer.

When any one of the 10 keys of the Binary Keyboard is operated a control voltage is obtained which is independently controlled by a potentiometer on the front panel of the unit. Ten separate potentiometers on the front panel provide this control. The keyboard is designed so that when two or more notes are operated simultaneously either the average voltage is provided or the sum of the voltages is provided, depending on a selector switch. In the case of an oscillator there is a logarithmic relationship between the oscillator's frequency and the control voltage. Therefore adding a given control voltage means the frequency of the oscillator is increased by a given increment. Similarly, averaging two voltages means the frequency obtained is the geometric mean of the origional frequencies. The keyboard control is arranged so that 4 keys can operate in either the summing or averaging mode, and the remaining 6 keys can be independently set to either the summing or averaging mode. In this way the keyboard can be used to obtain large frequency translations as well as very small changes.

The detection of the operation of a key was made using the change in electrical resistance between two insulated copper strips as it is touched. The copper strips were made using printed circuit board. It was found that the on resistance varied between wide limits and in some cases exceeded 1 MΩ. Furthermore it was found the 50 Hz pulses were present because of electric fields from the mains supply. The final method used to detect the actuation of a key was to use a field effect transistor in conjunction with a comparator. The field effect transistor provided extremely high input impedences, and the TTL output of the comparator provided immunity to the 50 Hz pickup. When capacitively loaded the output will charge quickly but will discharge relatively slowly. If the discharge time is made to exceed 20 ms then pulses at 50 Hz will sustain the correct output. It was found that there were three or so pulses until a firmer electrical contact was made.

A second design requirement arises when no key is actuated. It the control voltage were always to return to the same datum level the control level would soon give a very monotonous sound. The alternative is to hold the last voltage before the key was released.

This is relatively simple to achieve when only one note is depressed. When two or more notes are released simultaneously it would be desired that the combined note is sustained. In practice one note will be released momentarily before the other and a rapid electronic system will adopt the last key to be released, even if the additional time is only a few milliseconds. It is necessary to decide how much longer one note must be held than another so that the stores control voltage corresponds to that individual note rather than the combination. The value used was .2 seconds. If two keys are released within .2 seconds than the sustained control voltage will correspond to the combination of the notes. If the time interval is greater than .2 seconds then the effect of the previous key will be discarded.

It was found that a digital memory store was more suitable than an analog system.[1] This is because of its intrinsic long term stability and the fact that only 10 bits of information are required. The technique used to obtain this required memory and delay characteristics is given in the circuit of Diagram 5.3.A. This circuit is such that the control voltage adjusts rapidly to the release of the key. It the remaining keys are then released within .2 seconds then the control voltage will then return to that of the origional combination. In this way a rapid response to the operation of the keys is obtained, yet allowance is made for the slight errors when trying to release keys simultaneously.

The analog section of the circuit diagram is given in Diagram 5.3.B.

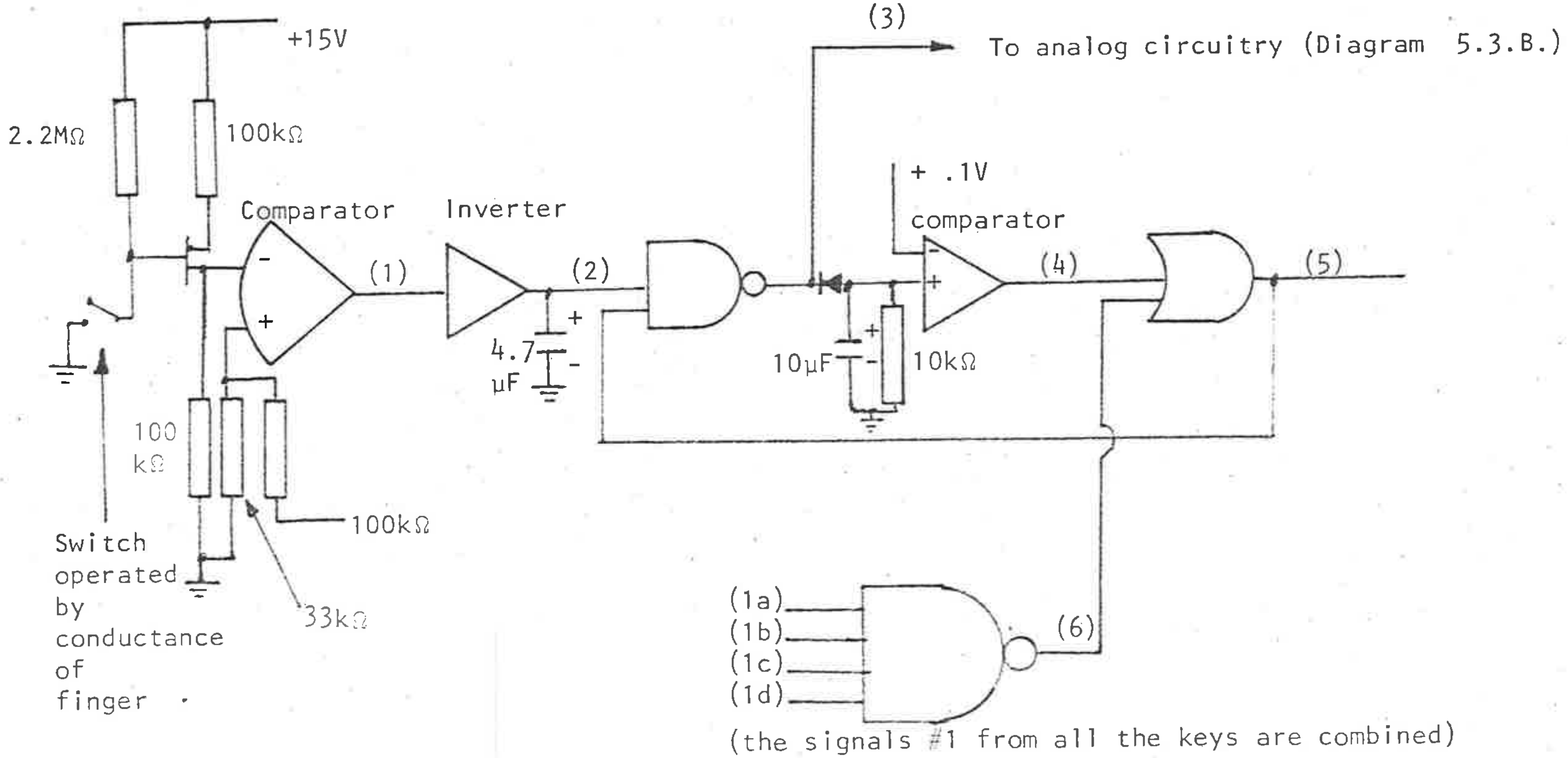


DIAGRAM 5.3.A. Digital section of the Binary Keyboard. One of these circuits is provided for each of the ten keys. See the next page for notes on the voltages.

Notes on the voltages in the circuit of Diagram 5.3.C.

(1)

Logic 1 when the key is actuated

Logic 0 when the key is not actuated

(2)

Logic 0 when the key is actuated

Logic 1 when the key is not actuated

The 0 → 1 transition is slow (20 ms)

The 1 → 0 transition is fast.

(3)

Logic 1 when the control voltage for the key is actuated

Logic 0 when the control voltage for the key is not actuated

(4)

Logic 0 when the control voltage for the key is actuated

Logic 1 when the control voltage for the key is not actuated with a .2 second delay for the 0 → 1 transition.

(5)

Logic 0 to hold the control voltage

Logic 1 to not hold the control voltage

(6)

Logic 0 when any key has been actuated within the last 20 ms

Logic 1 when no key has been activated within the last 20 ms.

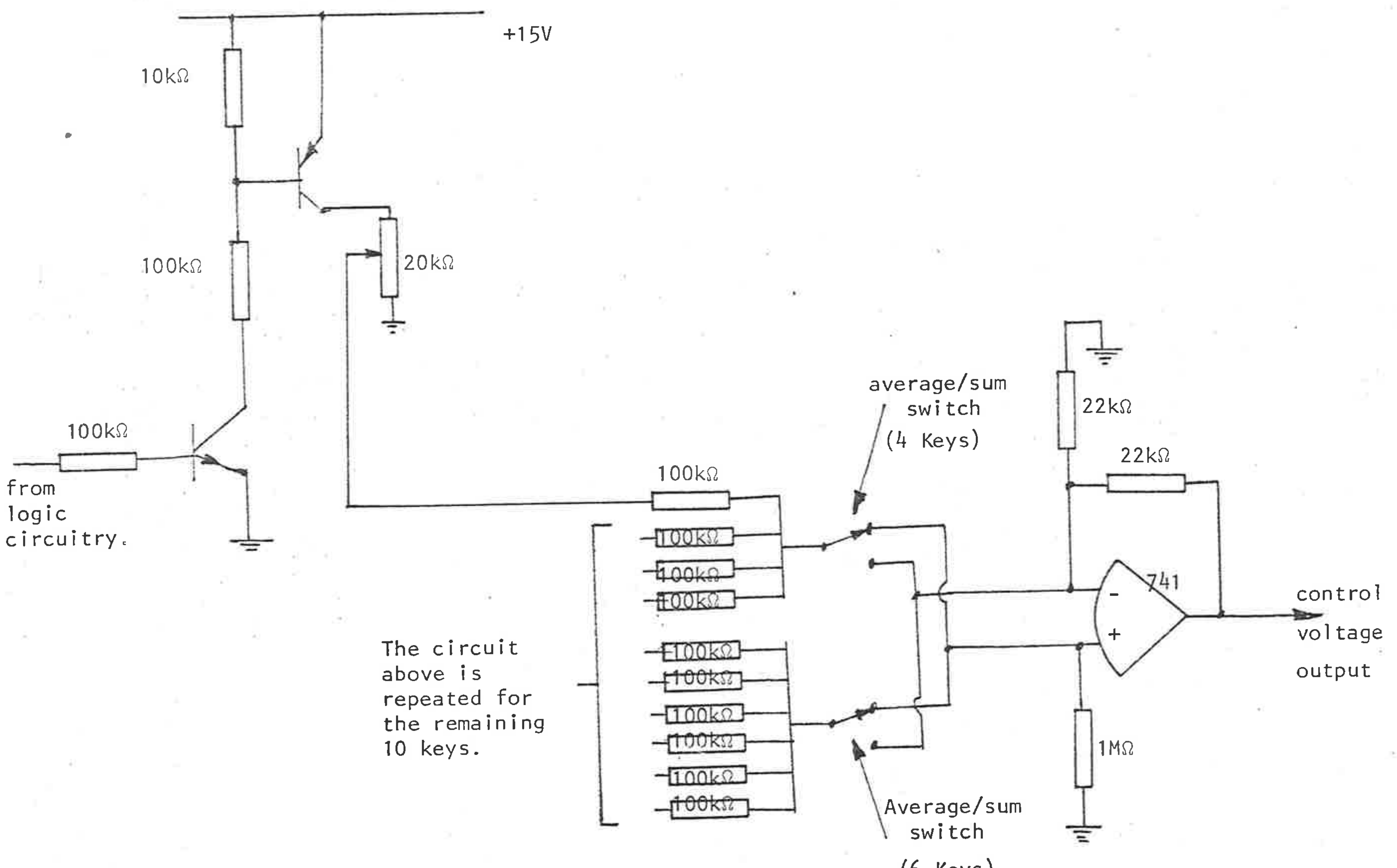


DIAGRAM 4.3.B. Analog section of the Binary Keyboard.

The pressure information from the Binary Keyboard is obtained using a pressure transducer which responds to very light pressures. Conventional transducers such as strain gauges were found to be far too insensitive. However Dynacon C variable resistance material is a pressure conductive polymer. The mechanical arrangement is given in Diagram 5.3.C., and it can be seen that an area of 4 cm by 8 cm was used under the 10 keys of the Binary keyboard. In this case the resistance to force relationship is as given on Diagram 5.3.D. The specification sheets of the material are given in appendix 1. The circuit diagram to obtain a control voltage which depends on pressure is given in Diagram 5.3.E.

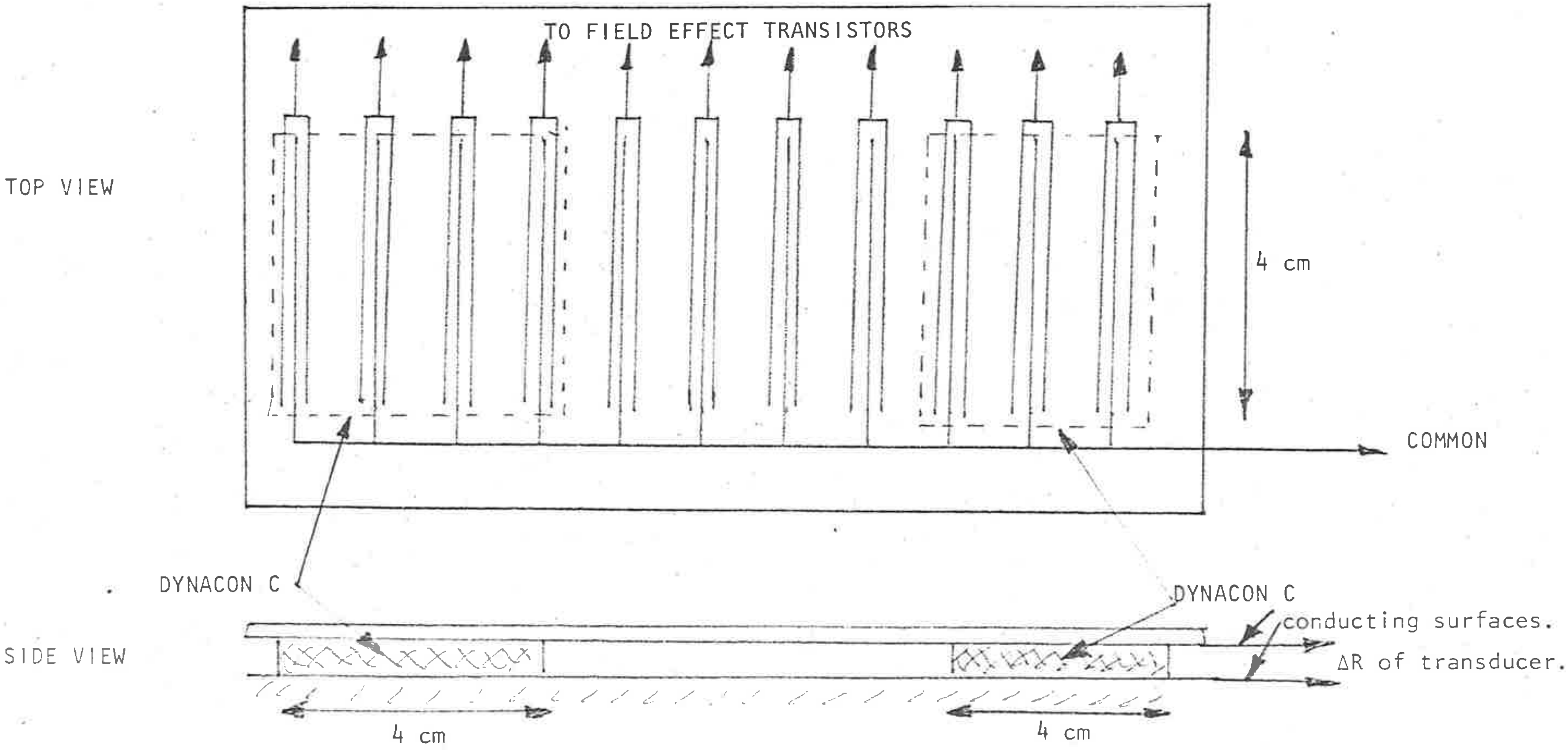


DIAGRAM 5.3.C. The mechanical arrangement of the Binary Keyboard showing the location of the DYNACON C pressure transducer. Two 4 cm square pieces are placed between conducting surfaces.

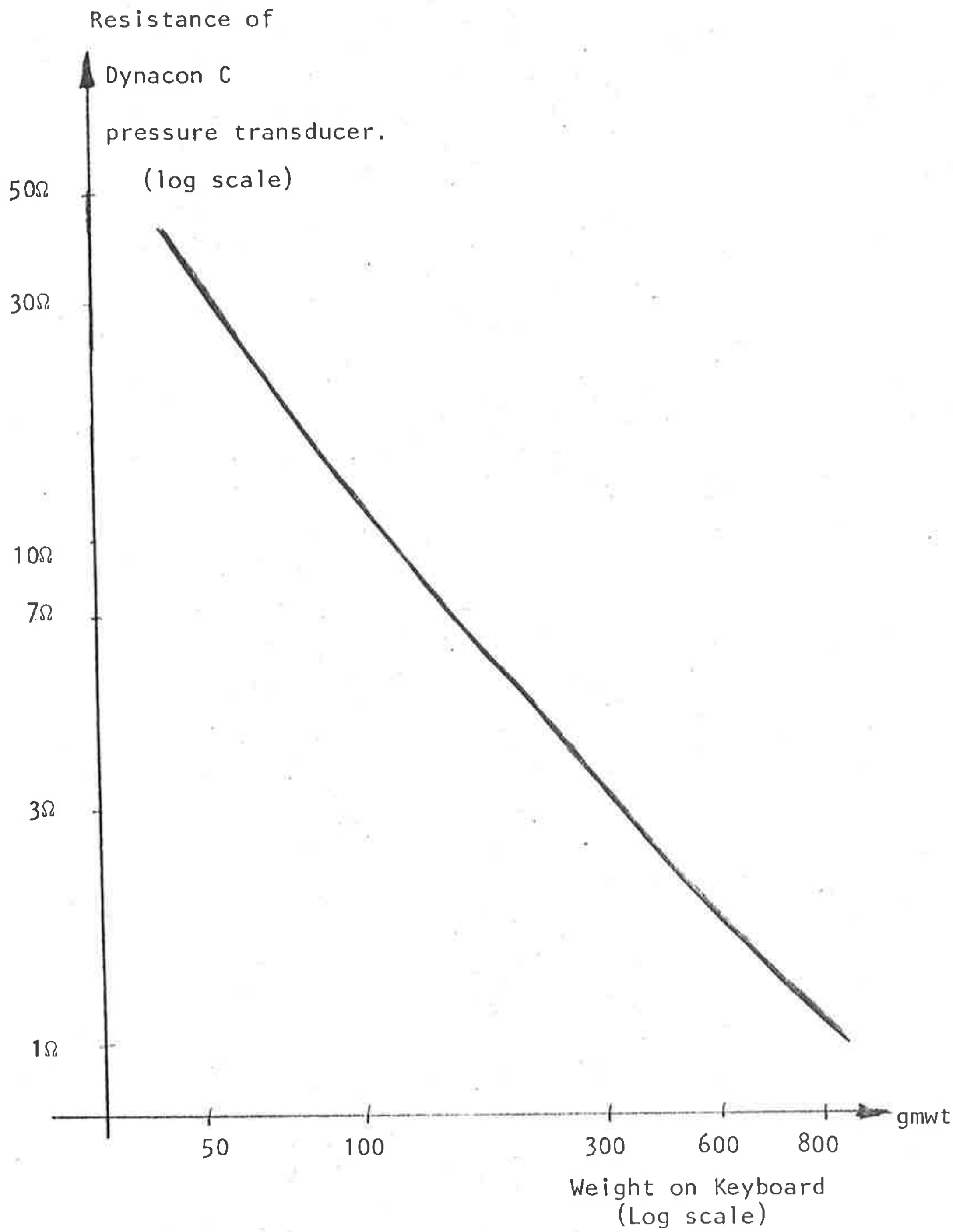


DIAGRAM 5.3.D. The resistance to force relationship for two 4 cm^2 pieces of Dynacon C pressure transducer as shown on Diagram 5.3.C.

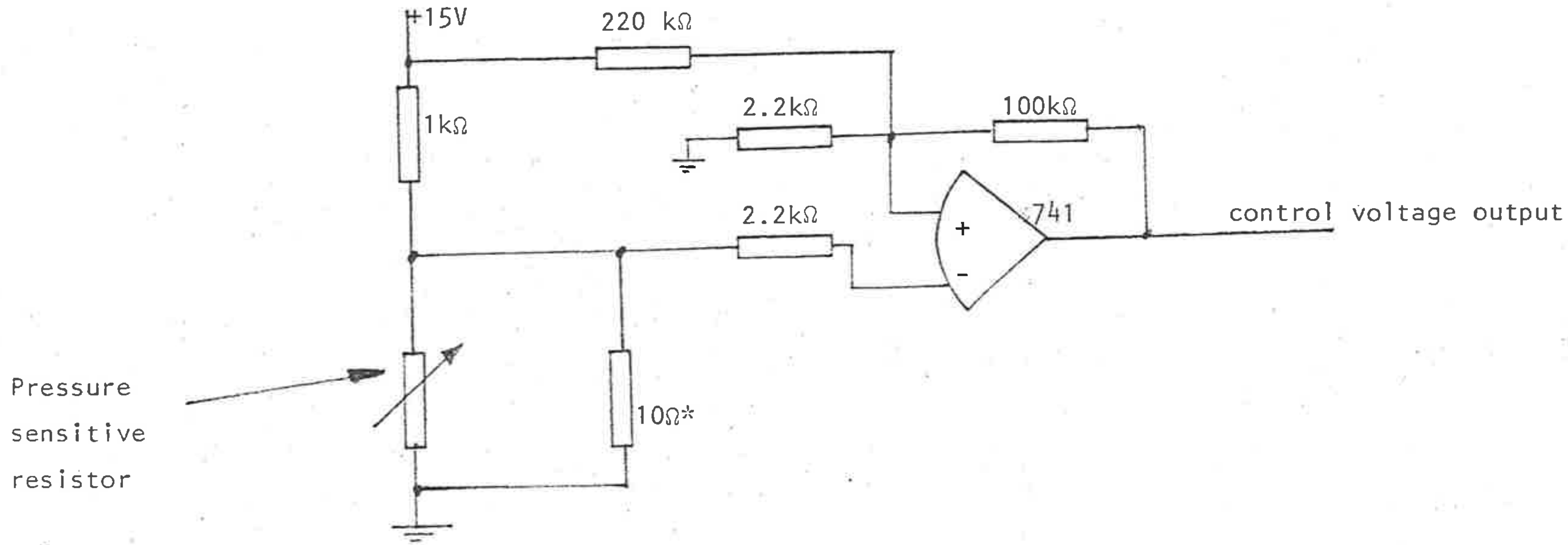


DIAGRAM 5.3.E. Circuit to obtain a control voltage from the pressure transducer. The resistor labelled * is chosen to give a suitable resistance range for the pressure sensitive resistor.

5.4 Lever Controller

With this controller the control voltage changes continuously with its physical movement. This is not true of the keyboards where the control voltage changes from one discrete voltage level to another. This means the sounds produced by this controller have characteristics which are distinct from those produced by the keyboards,even when the same sound sources are used.

With this controller a lever is attached to three orthogonal potentiometers as shown on Diagram 5.4.A. These provide three independent control voltage outputs. This is an extension of the slider potentiometer of the Moog synthesizer which has only one voltage output. With both systems the control voltage is held constant when the hand is released, but in the case of the lever controller this is achieved by mechanical friction rather than an electrical circuit. This means the oeprator can release the Lever Controller and then move it again without a discrete jump. This is not so with the electrical hold system of the slider controller because it is impossible to tell exactly where the controller was last touched. It does mean, though, the slider controller of the Moog synthesizer can provide both discrete jumps and a continuous control voltage.

The lever controller is similar to controlling systems which are used in very different applications. One example is the joystick of an aircraft. This similarity illustrates that many applications have an identical requirement; to control a process by a physical oepration. Although there are also very different requirements arising out of factors such as reliability and fatigue, the similarity does illustrate a useful engineering way of looking at a musical instrument.

The circuit to provide the range and the offset control for the voltage for the lever controller is given in Diagram 5.4.B.

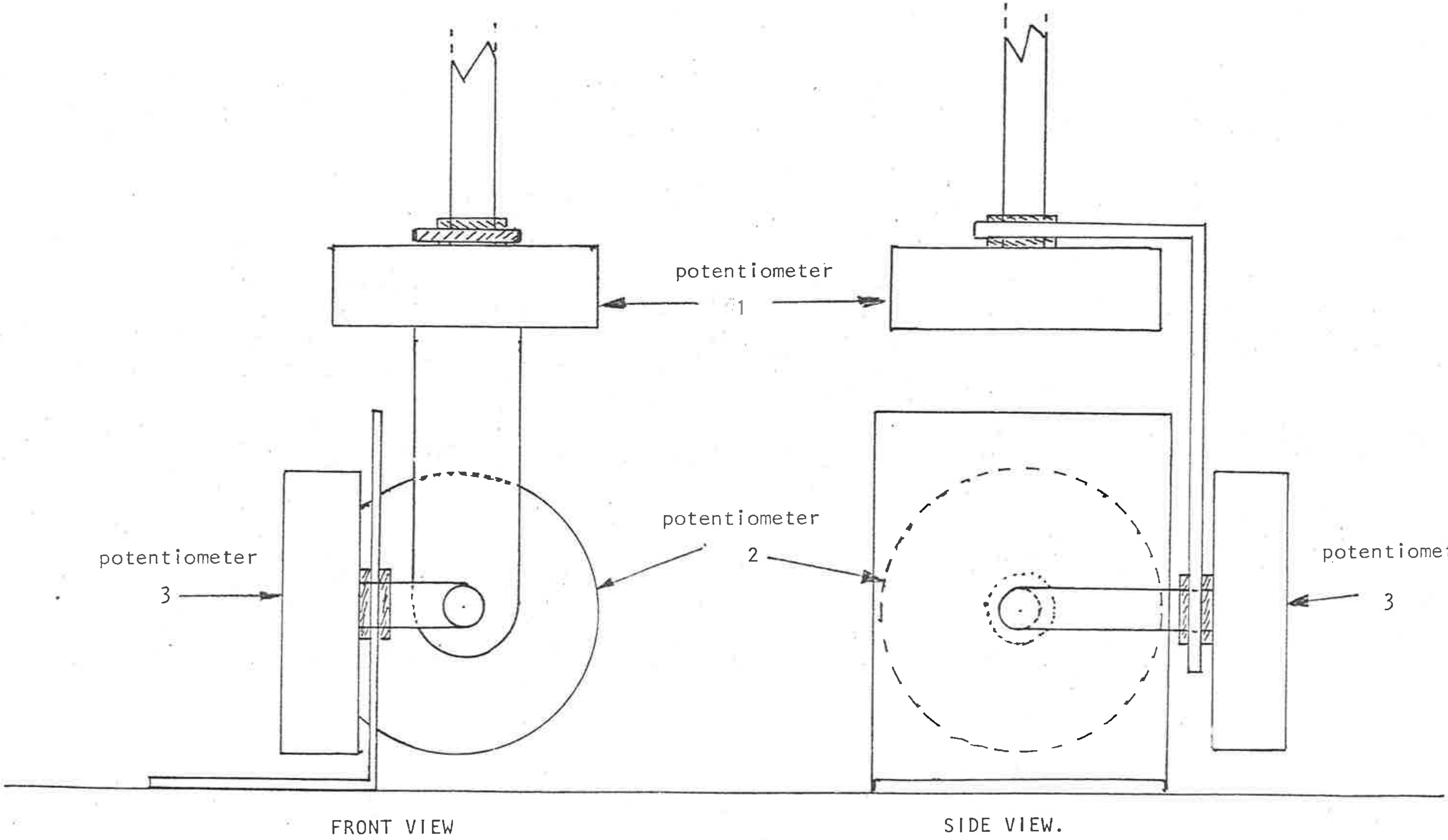


DIAGRAM 5.4.A. Mechanical arrangement of the potentiometers of the Lever Controller showing that the three potentiometers are placed orthogonally.

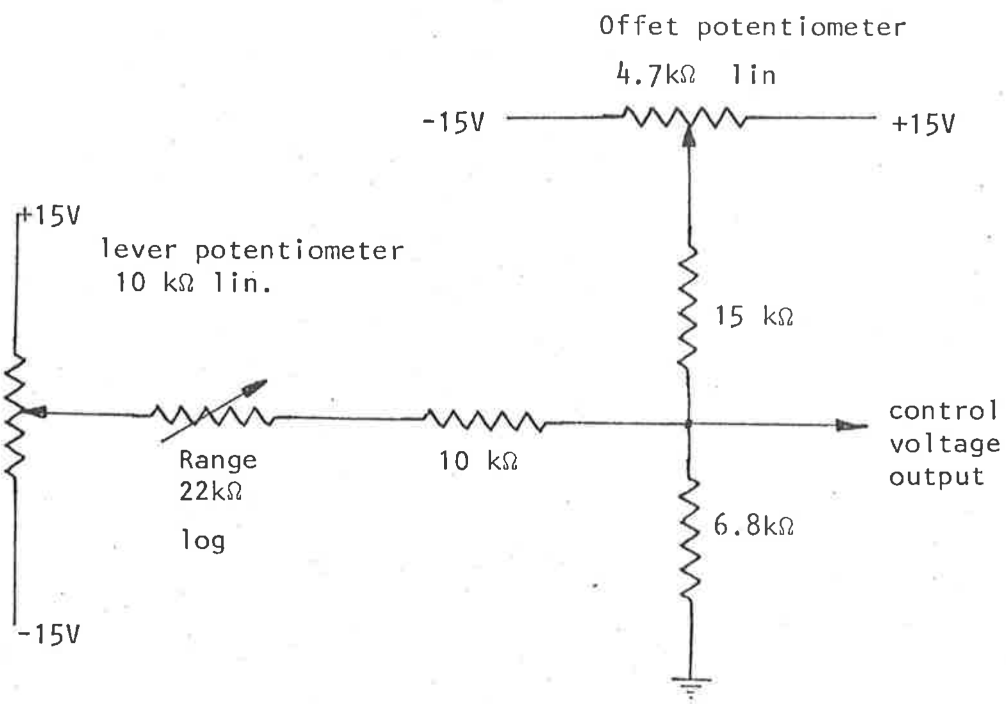


DIAGRAM 5.4.B. Circuit to provide the range and the offset voltage for the voltage from the lever potentiometers.

5.5 Foot Controller

A foot controller was designed to utilize other limbs. The ability to use the foot to control sound is important with many established musical instruments such as the piano-forte and organs. The mechanical arrangement is shown in Diagram 5.5.A. and the electrical circuit of Diagram 5.4.B. is suitable to provide the range and the offset voltages.

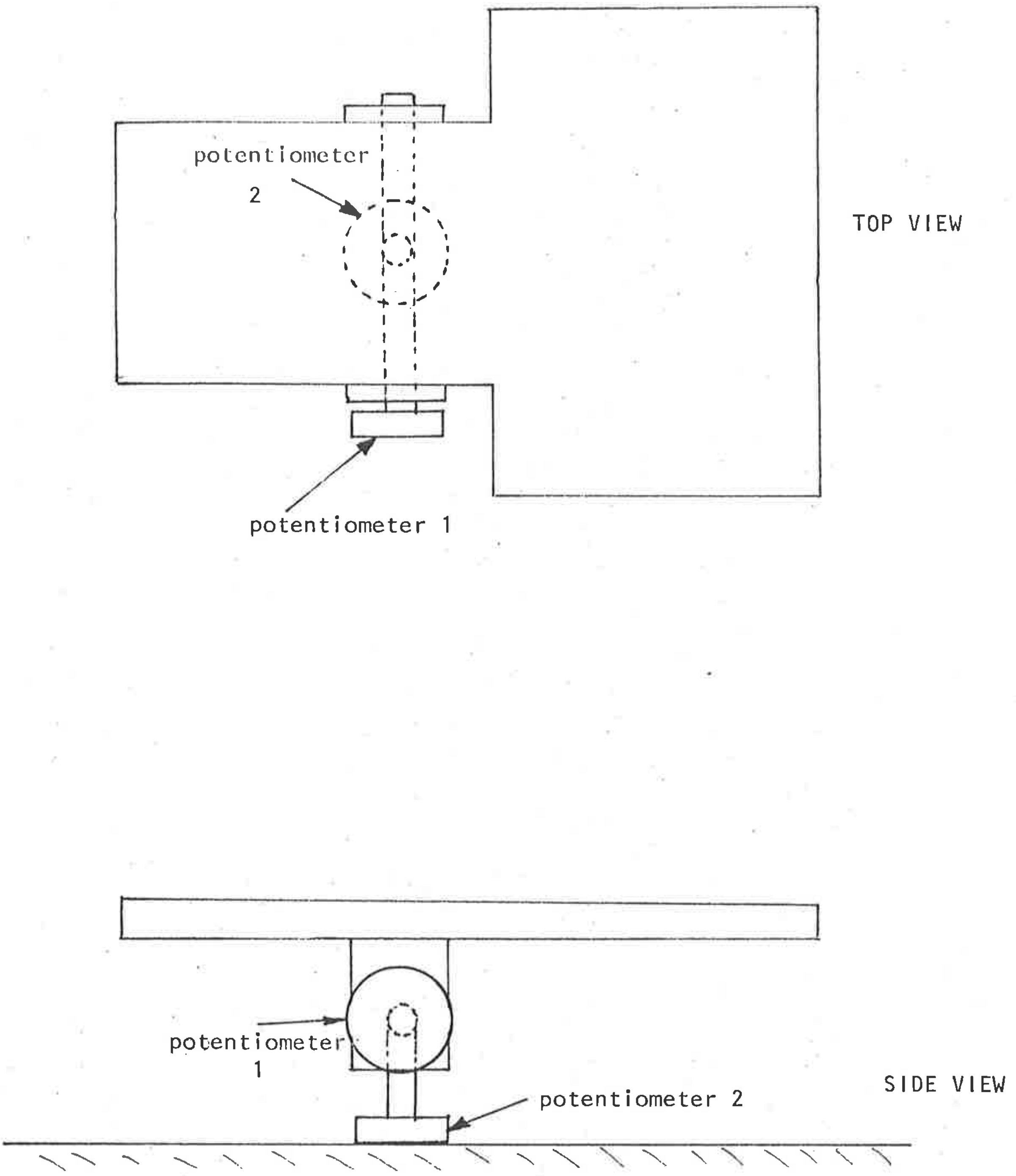


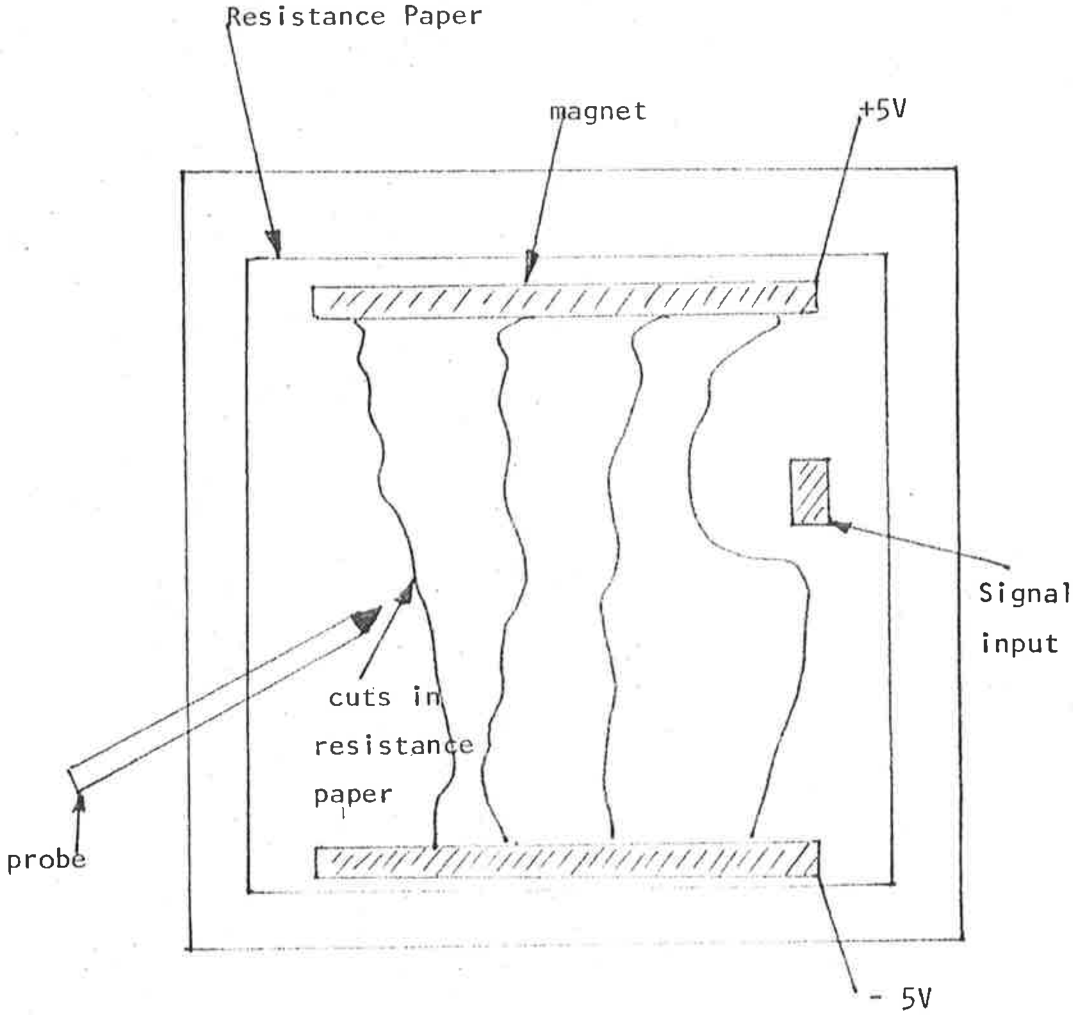
DIAGRAM 5.5.A. The mechanical arrangement of the potentiometers for the Foot Controller.

5.6 RESISTANCE BOARD CONTROLLER

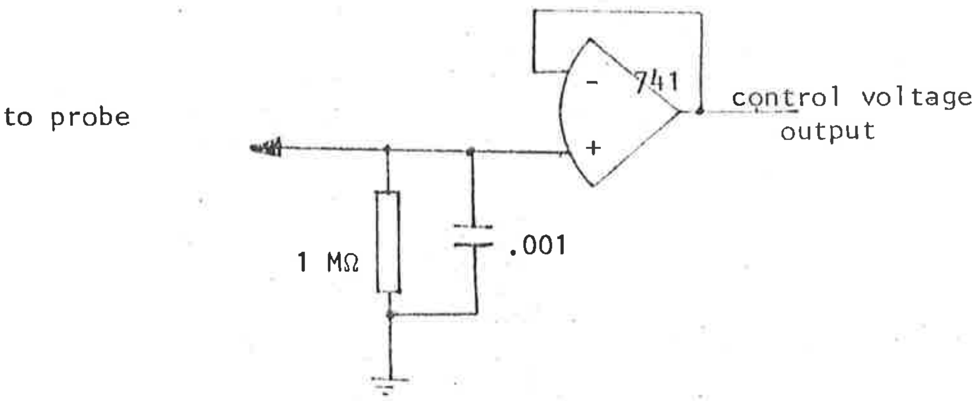
The conventional and binary keyboards (sections 5.2 and 5.3) give discrete jumps in the control voltage. The lever controller and the foot controller give continuous changes. The Resistance Board Controller is designed to give intermediate characterists; both discrete and continuous changes can be obtained on the same controller.

A probe is used to touch resistance paper which has a steady voltage set up across it. The distribution of the electric field on the paper depends upon the geometry of the paper. The paper can be cut to give various distributions, the thinner sections having the greatest potential gradients. Another way of increasing the flexibility of the controller is to introduce signal inputs into the paper's potential dirstibution. The use of magnets means the geometry of the Resistance Board Controller can be altered very conveniently.

The mechanical arrangement and the buffer of the resistance board controller are shown in Diagram 5.6.A.



Mechanical Arrangement



Buffer

DIAGRAM 5.6.A. Mechanical layout of the resistance board controller with its buffer circuit.

Magnets are used to locate the resistance paper.

5.7 Sound to Voltage Unit

5.7.1 Design Concept of the Sound to Voltage Unit

This unit illustrates the flexibility electronic techniques can provide in a sound synthesizer to control the sound produced. The generation and changing of signals by the synthesizer is controlled by voltages. In turn these control voltages can be from any suitable source. The sound to voltage unit is unique in that it produces control voltages from the amplitude and from the frequency of any input signal. If the voice is used as the source of such an input signal its extensive training can be utilized in ways which are not possible with conventional means of controlling musical instruments. The amplitude of the voice can be used to control the amplitude of quite dissimilar sounds: its frequency range can be translated, expanded, or even inverted.

5.7.2 The Amplitude Section of the Sound to Voltage Unit

The objective of this section is to provide a control voltage from the amplitude of an input signal. This can be achieved using a precision full wave rectifier which is followed by a low pass filter. Such a system is illustrated in Diagram 5.7.A., and typical waveforms for this system are given in Diagram 5.7.B. The average value of the magnitude of the voltage level is used to give the control voltage. This means a square wave of one volt peak will have the same voltage output as a sinewave with an amplitude of $\pi/2$ volt peak. The subjective assessment of the amplitudes of these two signals would be that the amplitude of the square wave is much greater because of its harmonic components.

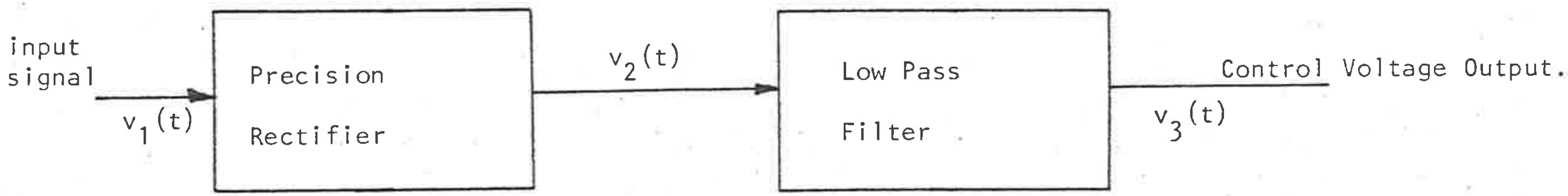


DIAGRAM 5.7.A. Block diagram of the simplified amplitude section of the sound to voltage unit.

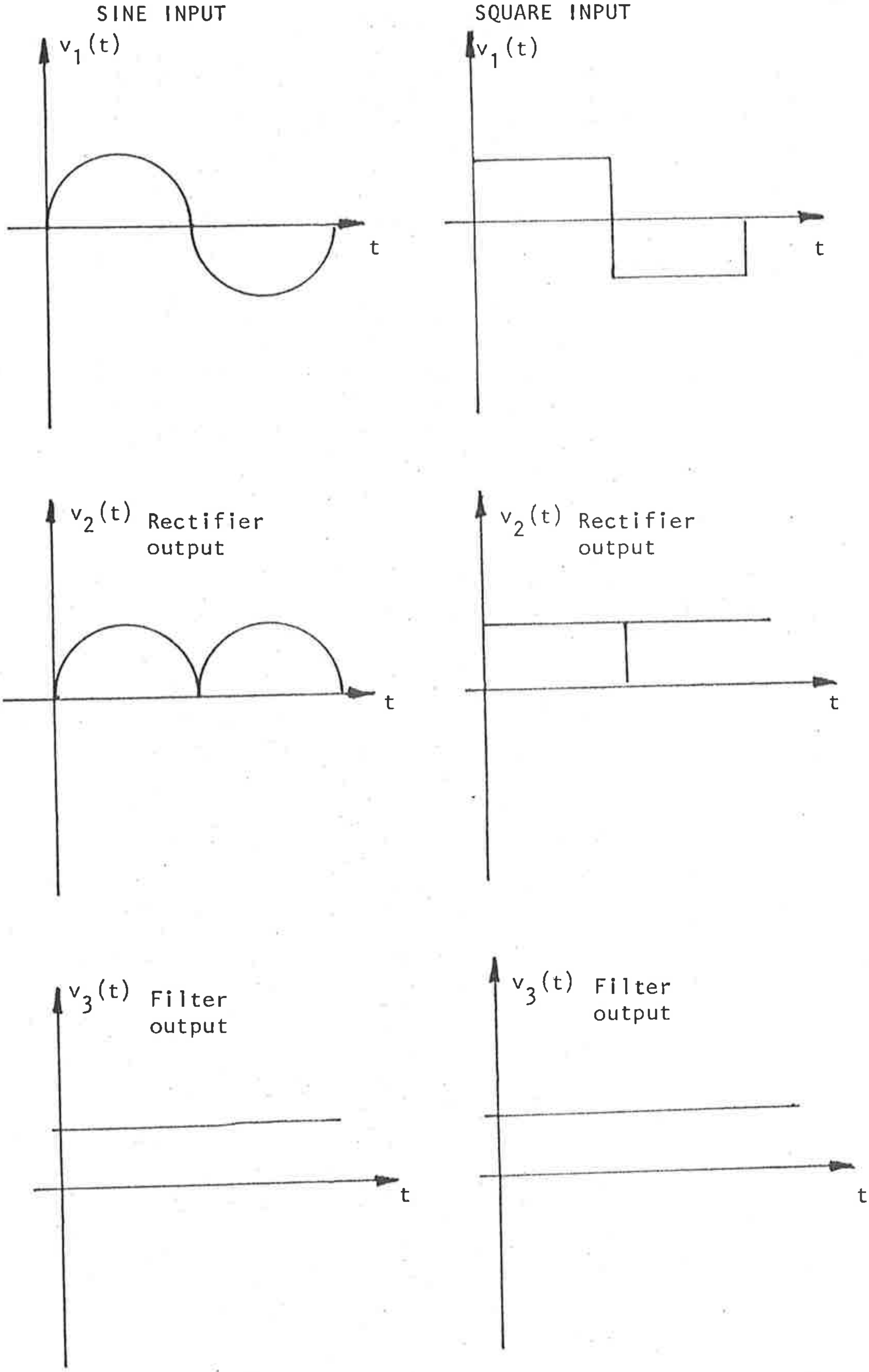


DIAGRAM 5.7.B. Typical Waveforms for the Block Diagram of Diagram 5.7.A.

Similarly a waveform approaching a train of impulses with the same control voltage output would be judged louder again. The effect of the harmonic components on the subjective assessment of the amplitude can be allowed for by passing the input signal through a tone control system which can be used to accentuate the frequency components of the waveform.

The ripple components from the signal rectifier is filtered using a third order low pass Butterworth filter which has a cutoff frequency of 5 Hz. This filtered output is passed through a diode function generator to achieve an approximately logarithmic relationship between the output voltage and the amplitude of the input signal.

Diagram 5.7.C. gives the complete block diagram of the amplitude section of the sound to voltage unit, and the circuit is given in diagram 5.7.D.

5.7.3. The Frequency Section of the Sound to Voltage Unit.

The objective of the frequency section of the sound to voltage unit is to obtain a control voltage which is proportional to the logarithm of an input frequency. In this way doubling the input frequency increases the output control voltage by a fixed voltage increment, and this complements the frequency characteristics of the sound sources.

A simple system to give this characteristic is outlined in Diagram 5.7.E. A zero crossing detector operates an integrator. The average output from the integrator is passed through a low pass filter and

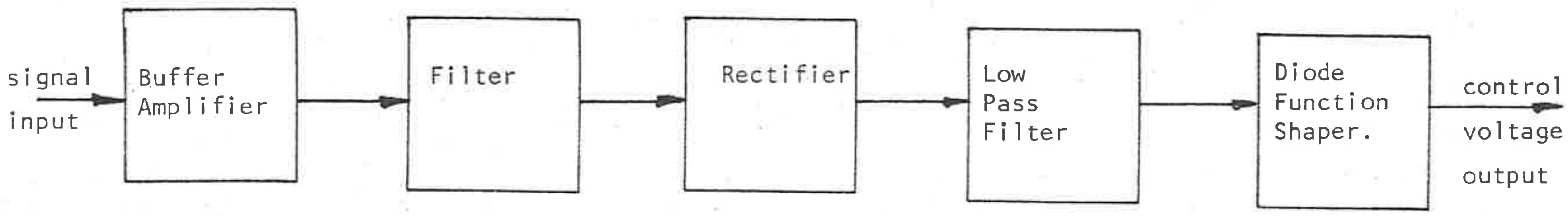


DIAGRAM 5.7.C. Complete Block Diagram of the Amplitude Seciton of the Sound to VOltage Unit.

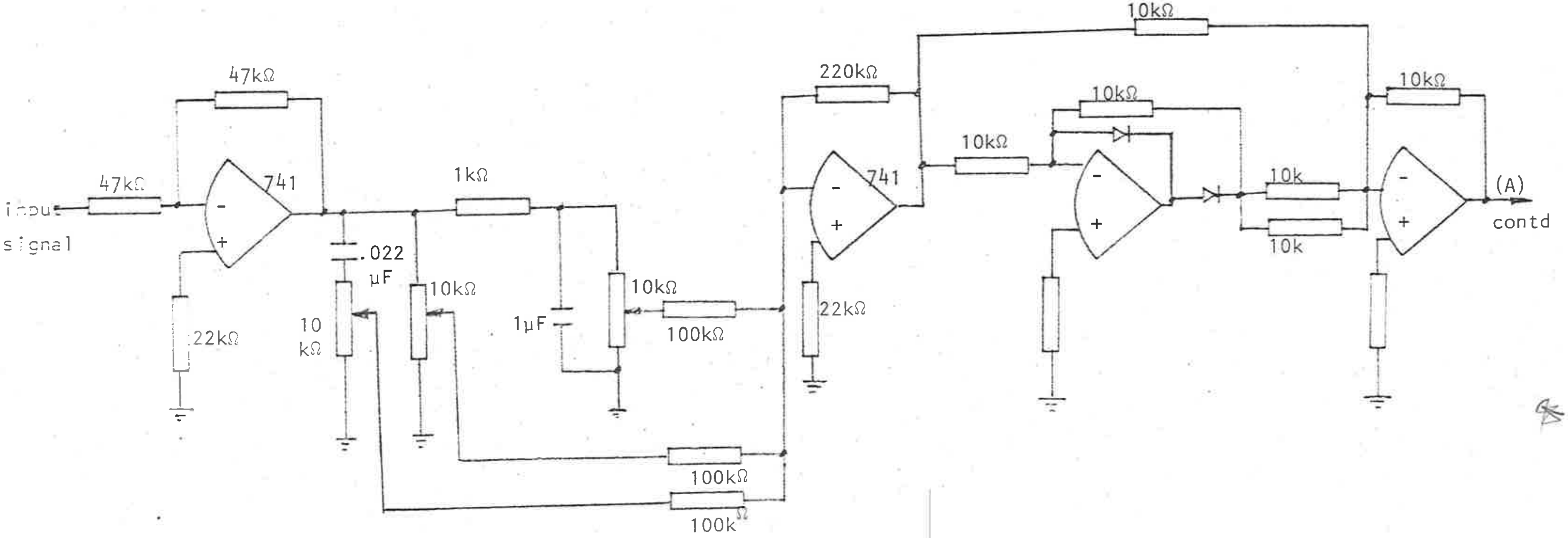


DIAGRAM 5.7.D. Circuit diagram of the sound to voltage unit. (Continued next page)

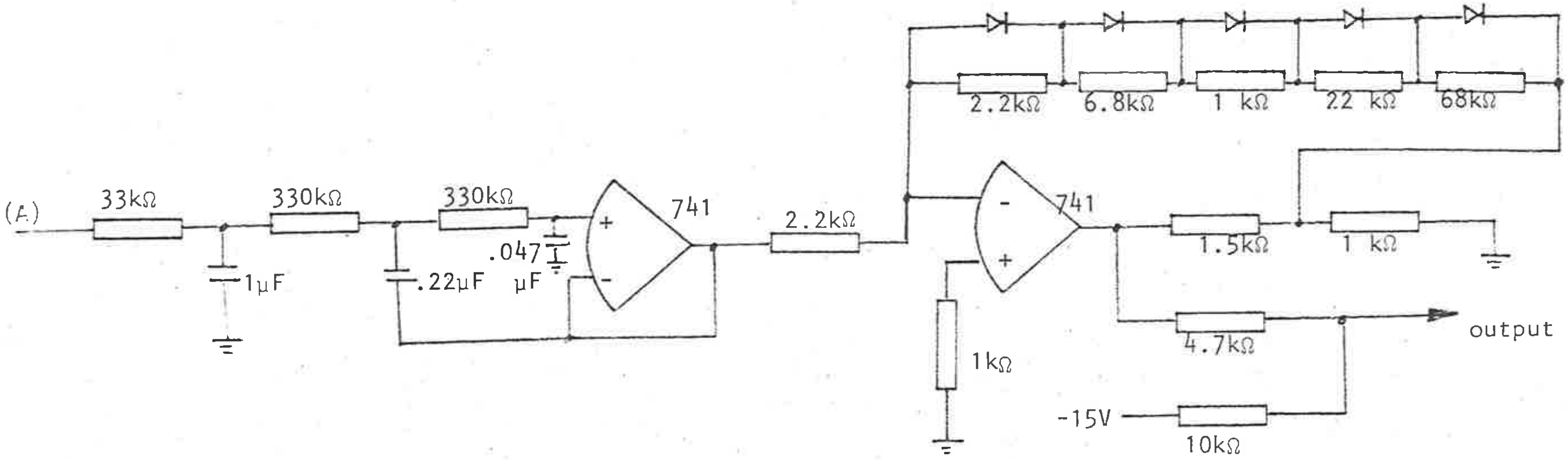


DIAGRAM 5.7.D. (Contd). Filter and shaper section of the amplitude section of the sound to voltage unit.

NOTE: A logarithmic voltage transfer relationship can be obtained using a diode or transistor junction characteristic. However this would introduce temperature dependence. Temperature compensated logarithmic ~~switches~~ converters are now available as integrated circuits.

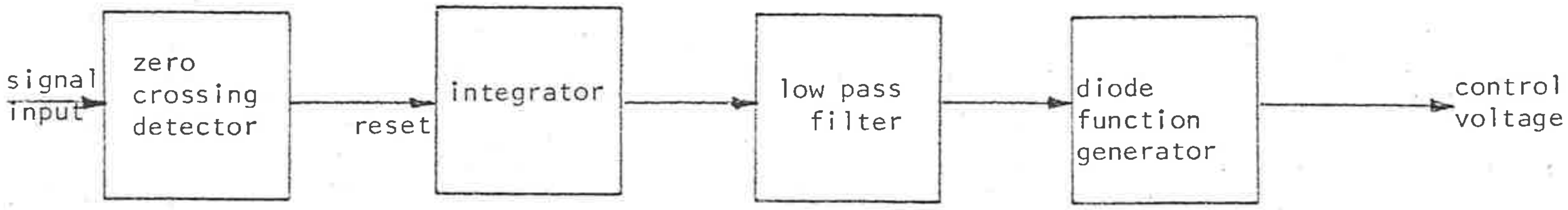


DIAGRAM 5.7.E. A simple system to give a control voltage from the frequency of an input signal.

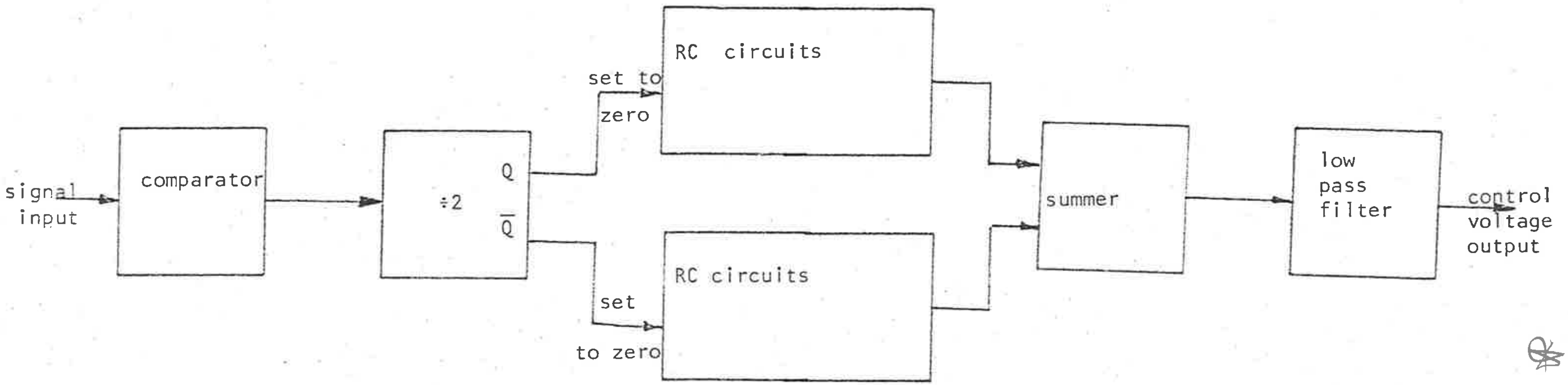


DIAGRAM 5.7.F. Block Diagram of the Frequency Section of the Sound to Voltage Unit

then a diode function generator is used to give a logarithmic frequency to voltage repationship. This simple system requires several requirements. The ratio of the input frequencies is $1 : 10^4$, and with this simple system the integration rate is constant. This means the average voltage from the integrator is also in the ratio of $1 : 10^4$. Since the maximum average voltage is 5 V, the highest frequency will give an average voltage of .5 mV. At this voltage level offset voltages are significant.

It is possible to avoid using such low voltage levels and at the same time perform the logarithmic frequency to voltage transformation. This approach uses a number of exponential terms of differing timeconstants. This design concept is also used in the Analog Frequency Standard (6.2) . The objective of this approach is to use the fact that although the frequency ratio is greater than $1 : 10^4$, never does the logarithmically related control voltage exceed $1 : 10^2$. Therefore the desired output voltage will not be subjected to errors from voltage offsets if it is generated directly. This design concept provides a basis for many useful applications including medium accuracy frequency meters (as described in section 6.2.), and oscillators.

The block diagram of the frequency section of the sound to voltage unit is given in Diagram 5.7.F. The input signal is divided by 2. This is because the input signals will, in general, have unequal mark to space ratios; the time the waveform is positive will differ from the time the waveform is negative. The division by two ensures that the intervals which are used to operate the R.C. lag circuits equal one complete period of the input signal if the input signal is periodic and does not have multiple zero crossings.

The exponential voltage across the capacitors in the R.C. sections are shown in diagram 5.7.G. The desired logarithmic period to voltage relationship is shown in the same diagram for comparison. It is desired that the average value of the sums of these voltages on the R.C. sections approximates the logarithmic relationship. The objective is to choose the best values of the timeconstants and asymptotic voltages to achieve the best approximation. The characteristice of the two waveforms differ at extreme values of time. As time approaches zero the logarithmic relationship approaches minus infinity whilst the exponential curves approach zero. As time approaches infinity the logarithmic relationship also approaches infinity whilst the exponential approaches a steady state value. It can be seen however that for intermediate times the two curves are of a similar form. The objective is to choose a series of exponential curves so that their sum is an optimum approximation to the logarithmic relationship over a given frequency range.

There are two parameters of each exponential curve which must be determined to give this approximation to the logarithmic relationship. These are the steady state voltage, V_{ss}, which the voltage across the capacitor will asymptote to as the time interval approaches infinity, and the timeconstant, T. For the charging RC circuit of diagram 5.7.H:

$v(t) = V_{ss}(1 - e^{-t/T})$ where $v(t)$ is the voltage across the capacitor at time t after the voltage was reset to zero.

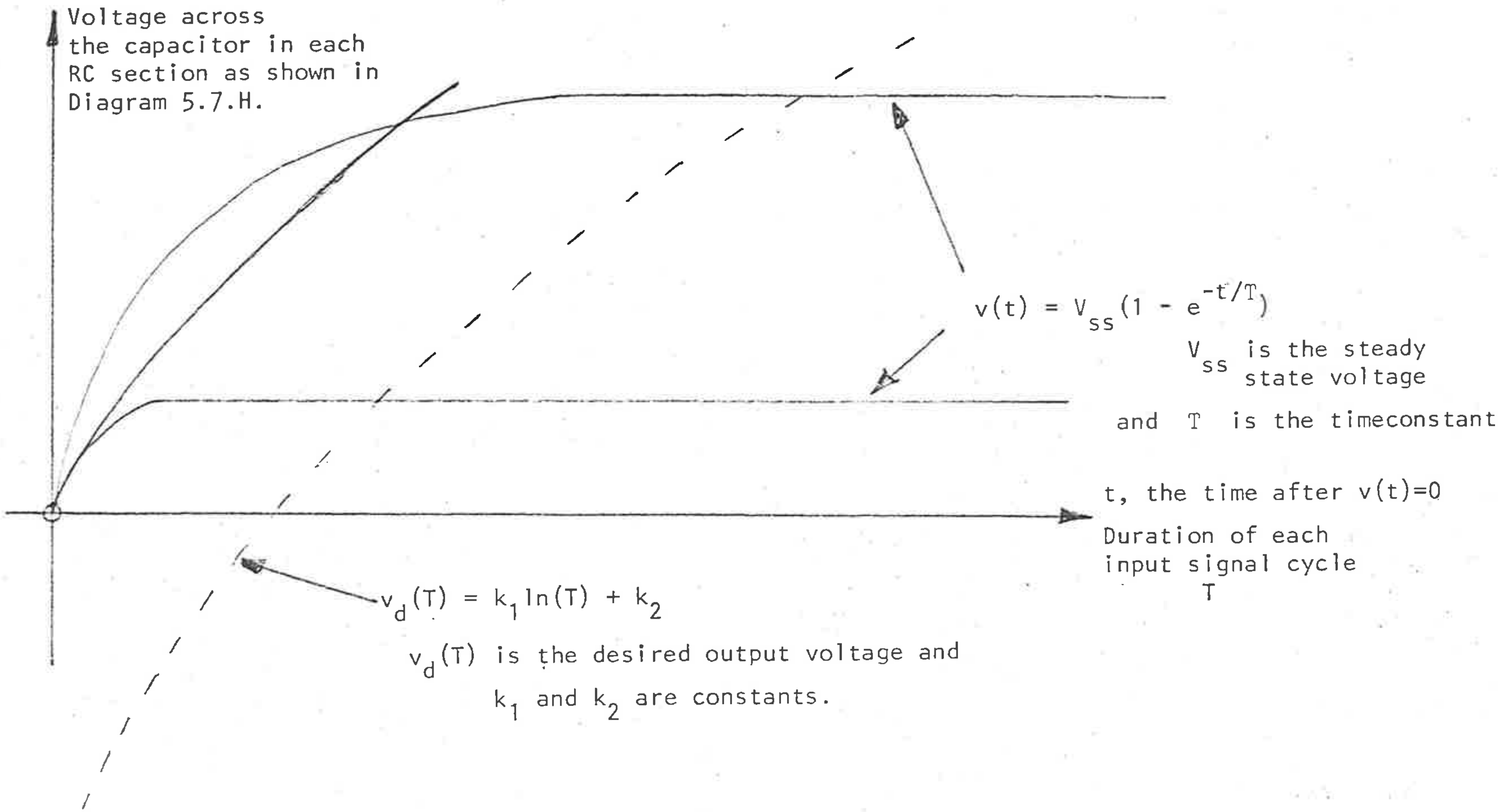
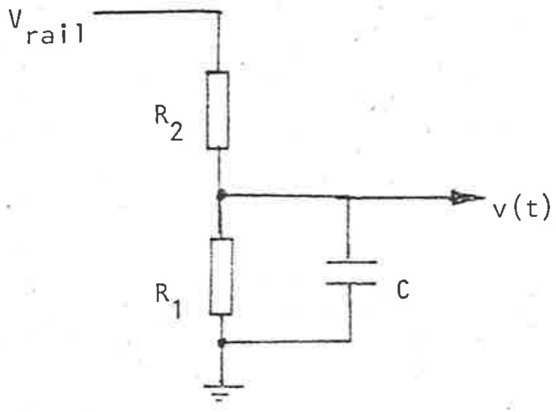


DIAGRAM 5.7.G. The exponential voltage across the capccitors in the RC sections of Diagram 5.7.F. The desired logarithmic relationship is also shown.



$$v(t) = V_{ss}(1 - e^{-t/T}), \text{ where } v(t) = 0 \text{ at } t = 0.$$

$$V_{ss} = V_{rail} \cdot R_1/(R_1 + R_2)$$

$$T = C/(1/R_1 + 1/R_2)$$

DIAGRAM 5.7.H. The simple RC circuit used to obtain the voltage function v(T) where T depends upon the time the circuit is allowed to charge to the setady state voltage level, V_{ss}.

The average voltage which is present at the output of the low pass filter which has the periodic input voltage with this waveform can be found for the case when the capacitor is shunted by the analog gate for a period T which depends upon the input frequency and for an equal interval, T, the capacitor is allowed to charge toward the steady state voltage V_{ss}.

$$v_{av}(T) = \frac{1}{2T}\int_0^T v(t)dt \quad , \text{ where } v_{av}(T) \text{ is the average voltage.}$$

$$= \frac{1}{2T}\int_0^T V(1 - e^{-t/\mathtt{T}})dt$$

$$= \frac{1}{2T}\Big[t - \mathtt{T}e^{-t/\mathtt{T}}\Big]_0^T$$

$$= \frac{V}{2}\Big[1 - \frac{\mathtt{T}}{T}(1 - e^{-T/\mathtt{T}})\Big]$$

The desired relationship is:

$$v_d(T) = k_1 \ln(T) + k_2 \quad , \text{ where } v_d(T) \text{ is the desired output voltage and } k_1 \text{ and } k_2 \text{ are constants.}$$

The constants k_1 and k_2 are arbitrary at this stage, and can be chosen to suit the intrinsic characteristics of the exponential functions. A linear voltage transposition can be made once the logarithmically dependent voltage is obtained.

The equation for $v_{av}(T)$ differs from the equation which arises in the case of the Analog Frequency Standard (6.2), but the technique which was used to solve for the optimum approximation is suitable for both applications

The desired relationship can be expressed as:

$$v_d(T) = k_1 u + k_2, \text{ where } u = \ln(T), \text{ ie } T = e^u.$$

u is a variable introduced so that:

$$\frac{dv_d(T)}{du} = k_1 \text{ , a constant.}$$

In this way the search for suitable parameters in the empirical relationship reduces to a search for parameters which give the best approximation to a constant value. For the empirical relationship,

$$\frac{dv_{av}(T)}{dT} = \frac{V}{2T}\left[\frac{\tau}{T}(1 - e^{-T/\tau}) - e^{-T/\tau}/T\right]$$

It follows:

$$\frac{dv_{av}(u)}{du} = \frac{V}{2}\left[\frac{T}{e^u}\left(1 - e^{-\frac{e^u}{\tau}}\right) - e^{-\frac{e^u}{\tau}}\right]$$

A small computer program was run to find the shape of the function $\frac{dv_{av}(u)}{du}$ as a function of u for a single RC section. The shape of this curve is given in Diagram 5.7.1, for the normalized case when $\tau = 1$ and $V = 1$. A consequence of this transformation of the frequency variable to u is that this basic trapeziodal shape is unaltered as the timeconstant or the steady state voltage is altered. Increasing the timeconstant gives a shift in the curve, and an increase in the steady state asymptotic voltage V increases

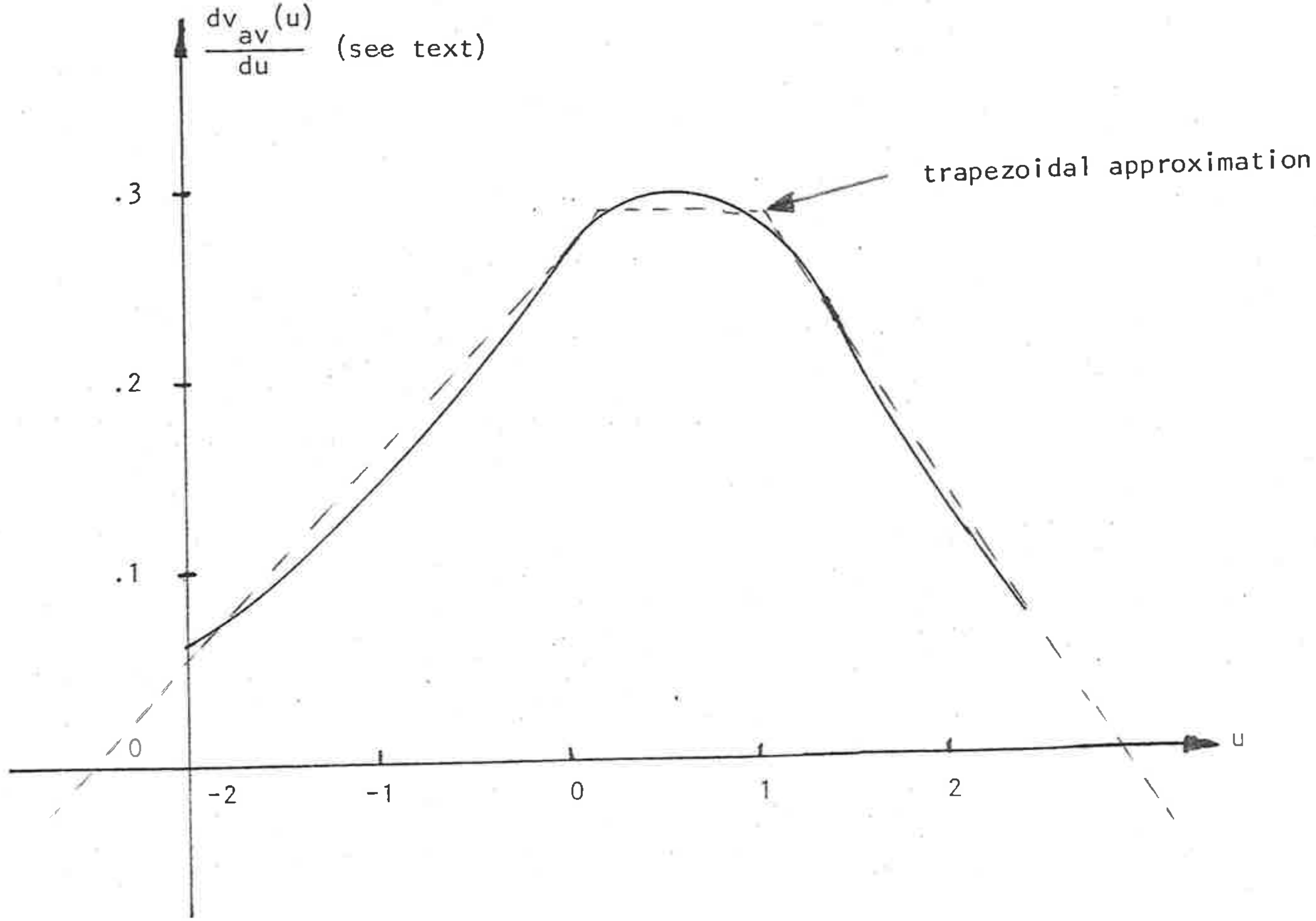


DIAGRAM 5.7.1. Plot of $\frac{dv_{av}(u)}{du}$ as a function of u. (see text).

the magnitude of the trapezoid in proportion. Knowing these properties the trapezoids can be combined so their sum gives a constant which corresponds to the logarithmic relationship. It can be seen that the trapezoids of equal spacing and of equal magnitude will give the best approximation to a constant. This corresponds to R.C. sections which have timeconstants which are in equal ratios to give the shift of the function of u by a constant increment to the right. In order to obtain the constant magnitude of the function of u the asymptotic voltage is held constant. This is shown in Diagram 5.7.J.

It may be thought that the more R.C. stages for a given frequency ratio the better the approximation. The use of trapezoids shows that this is not necessarily the case. A computer program calculated the error of the logarithmic approximation as a function of the ratio of the successive timeconstants. The graph of Diagram 5.7.K. shows that the error is ±.42% for successive ratios of 3.9 in the timeconstants of each RC section. If the trapezoids are put closer together than this then there is more error in the approximation as well as more hardware required. It is possible to use exactly twice or three times this optimum ratio to further reduce the error however.

A direct application of this principle is the measurement of frequency. A moving coil voltmeter on the buffered output of the RC stages would give a logarithmic scale for frequency. Frequency division of the input signal would give a very versatile instrument.

The circuit of the frequency section is given in Diagram 5.7.L. to Diagram 5.7.N.

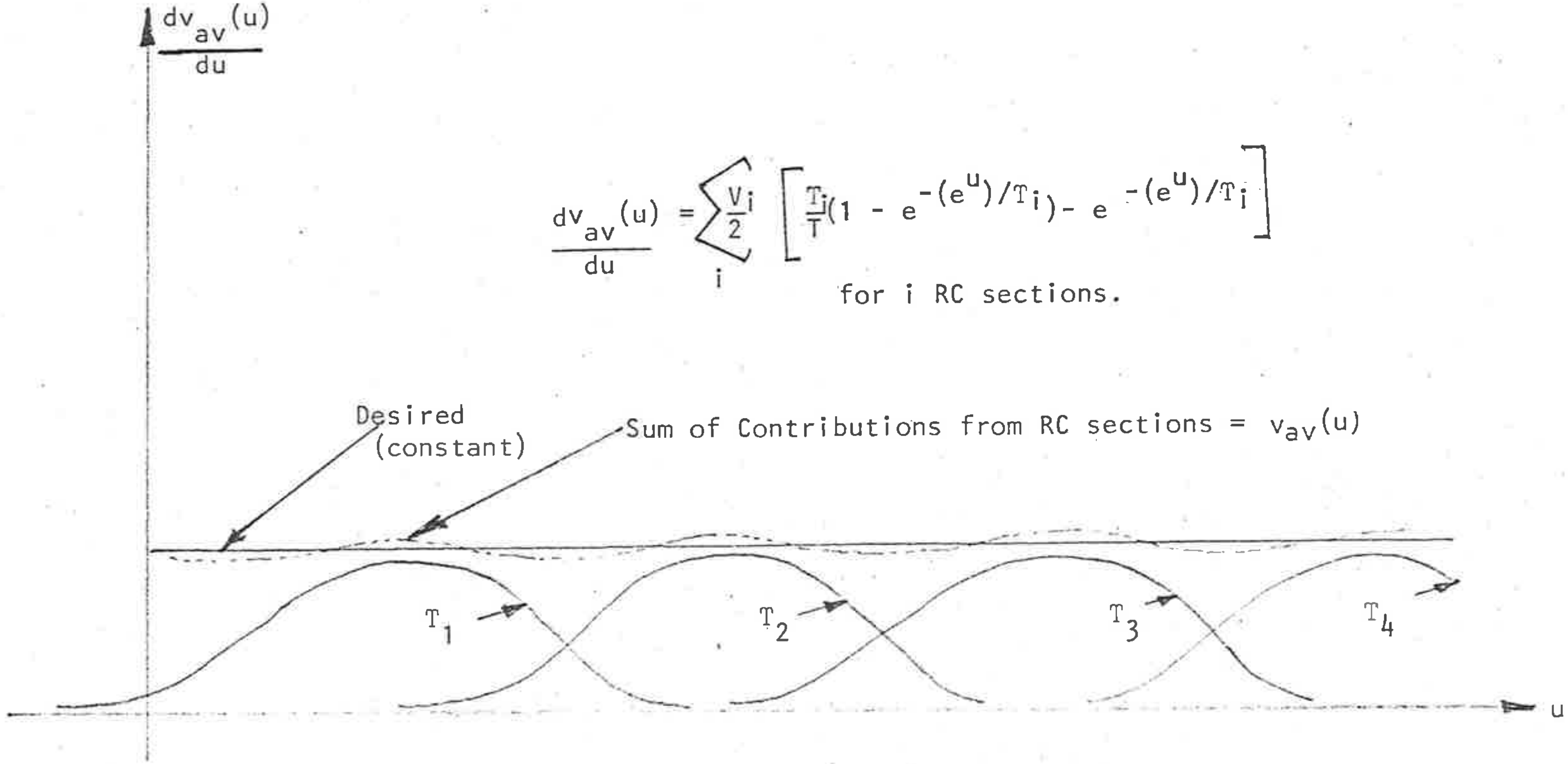


DIAGRAM 5.7.J. Plot of $v_{av}(u)$ as a function of u showing how the trapezoids corresponding to each timeconstant add to a curve which is nearly constant. (A constant value is desired).

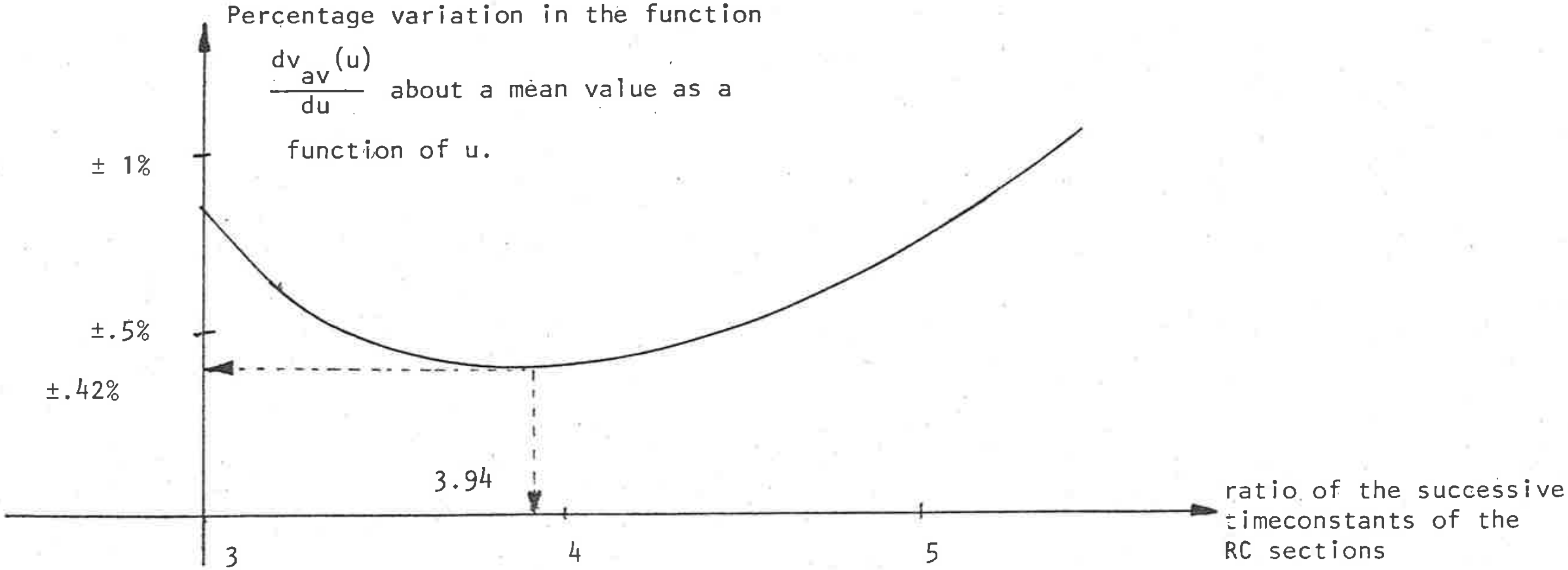


DIAGRAM 5.7.K. The percentage variation of the function $\frac{dv_{av}(u)}{du}$ about its mean value as a function of the ratios of the successive timeconstants.

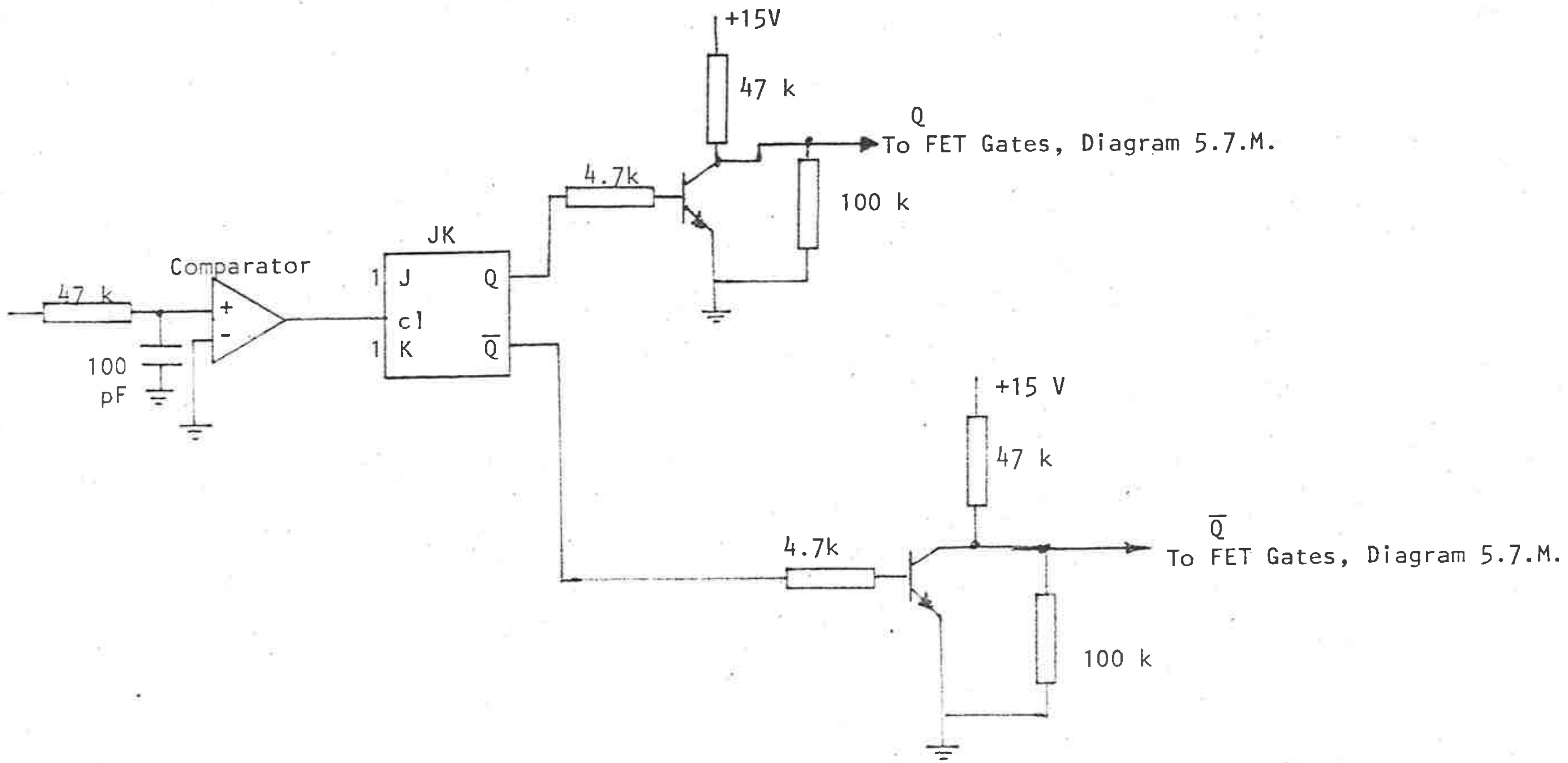


DIAGRAM 5.7.L. Comparator of the frequency section of the sound to voltage unit.

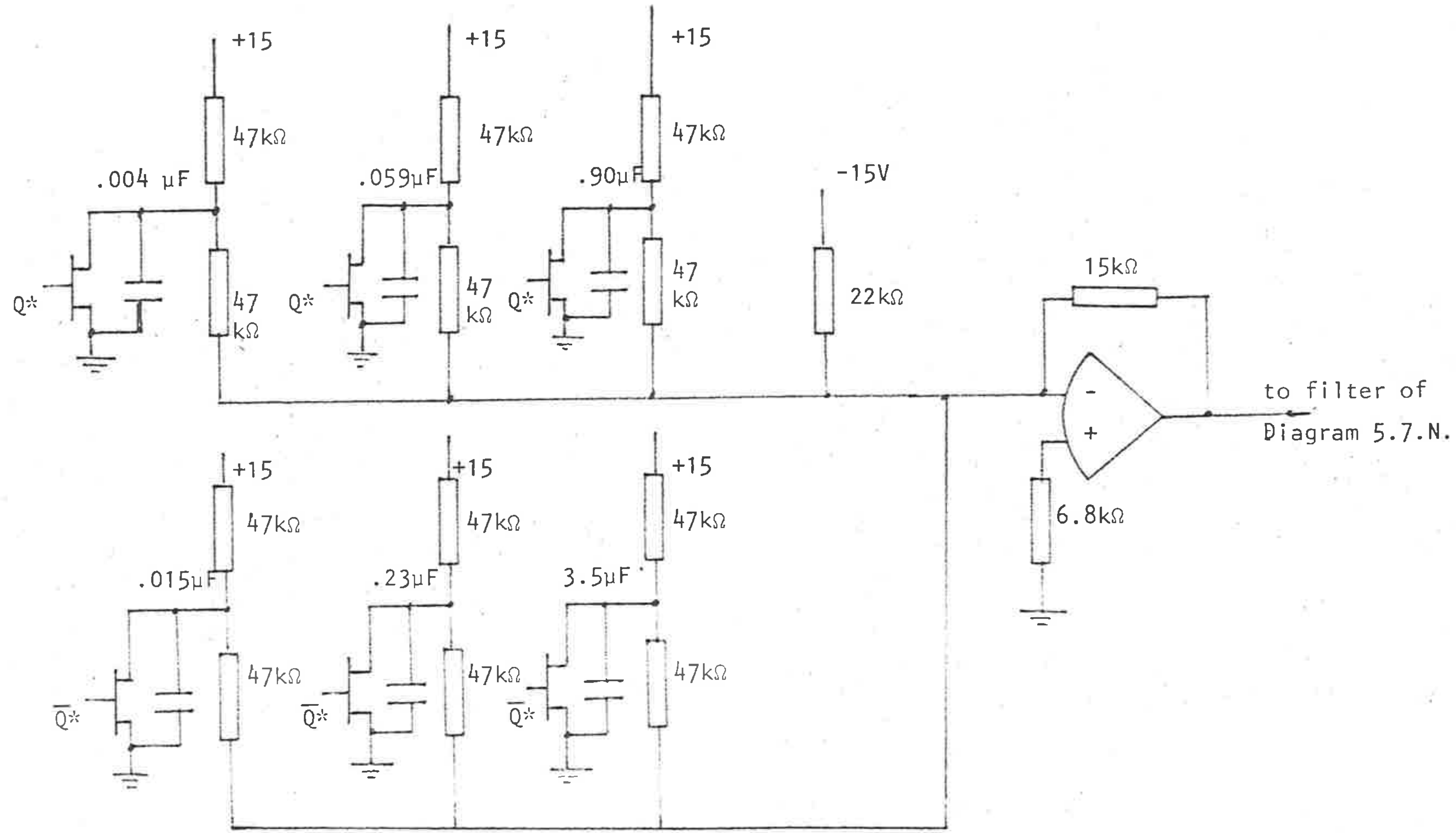


DIAGRAM 5.7.M. RC sections of the frequency section of the sound to voltage unit. (*See Diagram 5.7.L)

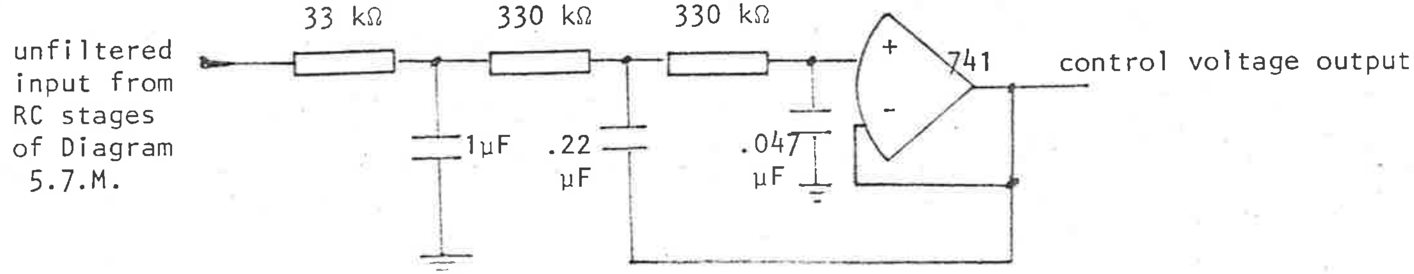


DIAGRAM 5.7.N. Filter section of the sound to voltage unit.

5.8 Control Voltage Manipulator

The control voltage manipulator takes advantage of the design philosophy of the music synthesizer by using the control voltage to change the relationship between the operator's physical movements and the corresponding sound of the synthesizer.
The manipulator provides integral, differential and proportional components of the input control voltage, and it simply a tone control system with very low cutoff frequencies, The circuit diagram of the control voltage manipulator is given in Diagram 5.8.A.

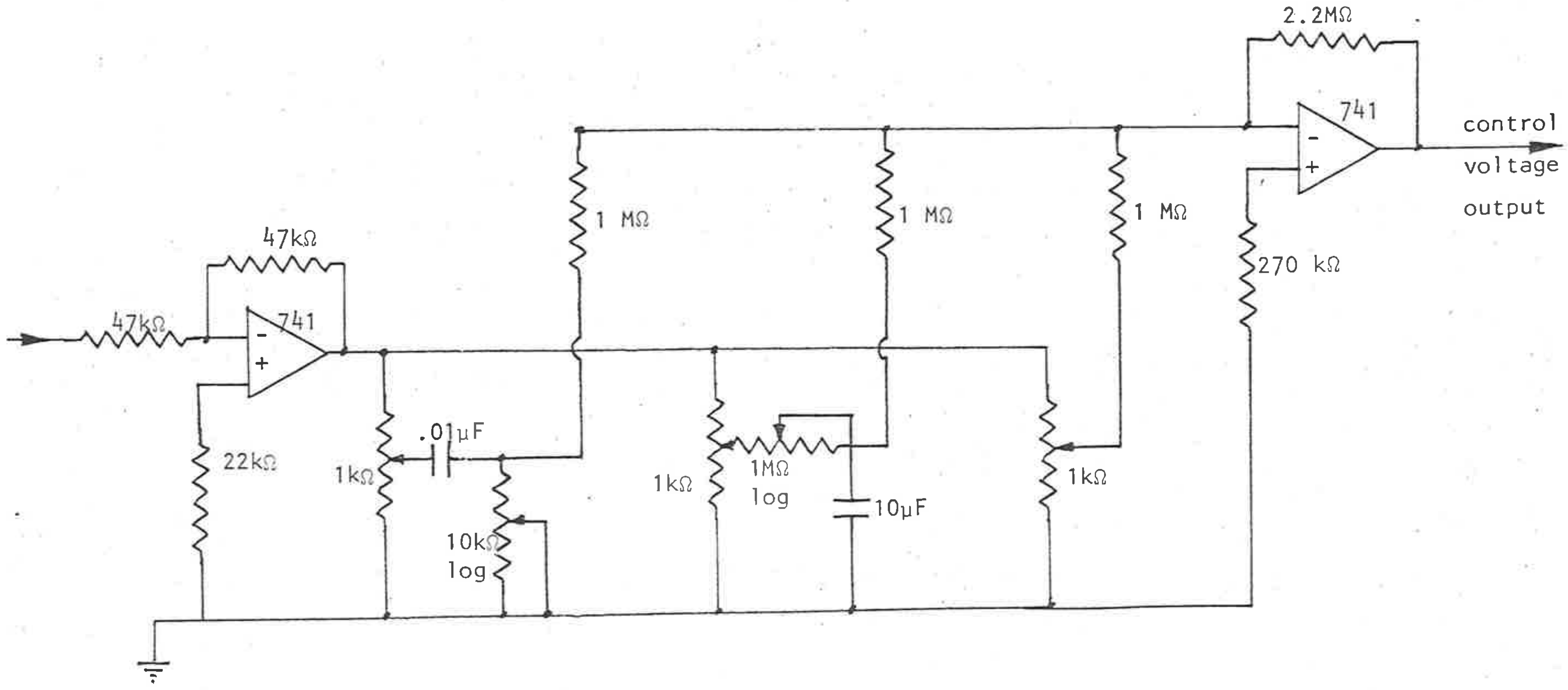


DIAGRAM 5.8.A. Circuit of the Control Voltage Manipulator.

NOTES FOR CHAPTER 5

1. Analog storage techniques have improved considerably in recent years. Voltage followers such as the LH 0033 have typical input bias currents of 1 nA maximum. See reference 19.
In this application the digital storage technique is still preferred because of its intrinsic long term stability.

CHAPTER 6 FREQUENCY STANDARDS

6.1 The Requirement for Frequency Standards

The requirement for frequency standards in an electronic music synthesizer depends upon the musician's compositional technique. If the sounds used are of very short duration or always involve a sweeping of the frequency, then accurate measurement of the frequency is not important. In the case of two sustained periodic waveforms the ear responds to the small frequency differences between the harmonic components of the two waveforms. In this case it is necessary to accurately control the ratios of the frequencies, .1 % being a typical requirement. This requirement is most stringent when the frequencies have considerable harmonic components which are in the middle range of the audible spectrum. A relatively large change in the absolute pitch of all frequency components can be tolerated by most listeners, a 1 % long term change being well within the acceptable range.

The requirements of a frequency standard follow from these stability requirements. The short term accuracy of the control units and oscillators of this synthesizer is adequate for most requirements. The design of oscillators and control units with a long term stability of .1 % would be very expensive. To achieve this resources would have to be diverted from other features of the synthesizer.

The frequency of a signal can be adjusted by comparing it with a reference signal such as a tuning fork. The more common engineering approach is to measure the frequency of the signal and to use this reading for the adjustment. The direct reading of frequency requires a digital readout to give sufficient accuracy, and until recently these were prohibitively expensive. The reference oscillators were designed to use the principle of a tuning fork by providing a reference signal to allow adjustment of the dependent signal.

Two oscillators were designed to give this reference frequency. The Analog Frequency Standard is suited to the tuning of even tempered scales where the frequencies of successive notes are in a geometric progression. The Hybrid Frequency Standard uses analog control over a limited range of frequencies. This oscillator uses a crystal oscillator as a frequency standard, and derives the dependent frequency using digital and analog techniques. The design of both oscillators is believed to be novel.

6.2 The Analog Reference Oscillator

6.2.1 Introduction

This oscillator has a logarithmic relationship between the control voltage and the frequency. It uses no nonlinear elements, apart from a voltage comparator which has accurately determined characteristics. This gives the oscillator good intrinsic frequency stability and accuracy.

The logarithmic control voltage to frequency relationship facilitates the setting up of even tempered scales.[1] Even tempered scales are sequences of notes with frequencies which are in a geometric progression. The corresponding control voltages are in an arithmetic progression when there is a logarithmic control voltage to frequency relationship.

Let $f = k_1 e^{k_2 v}$ where f is the frequency of the oscillator

v is the corresponding control voltage

and k_1 and k_2 are constants

Let $$\eta = f_2/f_1 = \frac{k_1 e^{k_2 v_2}}{k_1 e^{k_2 v_1}} = e^{k_2(v_2 - v_1)} = e^{k_2 \Delta v}$$

Therefore a frequency ratio of η corresponds to adding Δv to the control voltage of the oscillator, irrespective of the datum frequency f_1. It follows that if the control voltage is divided into equal increments then the corresponding frequencies will be in a geometric progression. There is a resistive divider provided in the oscillator with 6 resistors which are given binary weighting. This allows the division of an increment in the control voltage; an arithmetic progression with up to $2^6-1 = 63$ terms can be fitted between and two voltage levels. This means that the corresponding frequencies are in a geometric progression; it is possible to divide the interval between any two frequencies into a geometric progression which has up to 63 terms. This means the musician can tune even tempered scales without reference to mathematical tables.

6.2.2 The basic circuit

The basis of the design of the oscillator is given in the circuit in diagram 6.2.A. A comparator has an output of +V volts when the voltage on the noninverting input (v+) is greater than the voltage on the inverting input (v-). Otherwise the voltage on the output is -V volts. The noninverting input is connected through a potential divider to the output voltage such that v+ = kV, where V is the output voltage. The voltage on the inverting input, v-, is connected to the output voltage via a first order lag circuit.
If v+ > v - the output of the comparator will be positive. The capacitor on the noninverting input will charge until v- exceeds v+. Then the output will switch to -V volts, and the potential on the noninverting input will change simultaneously because of the divider action. Then the voltage on the capacitor will drop until v- is more negative than v+.

The output will then go to +V volts and the cycle will be repeated. The frequency of this oscillation depends upon 3 factors.

(1) The divider factor k which is present at the non-inverting input.

(2) The RC timeconstant,

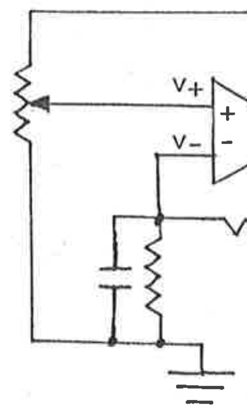


DIAGRAM 6.2.A Basic Circuit of the Analog Frequency Standard

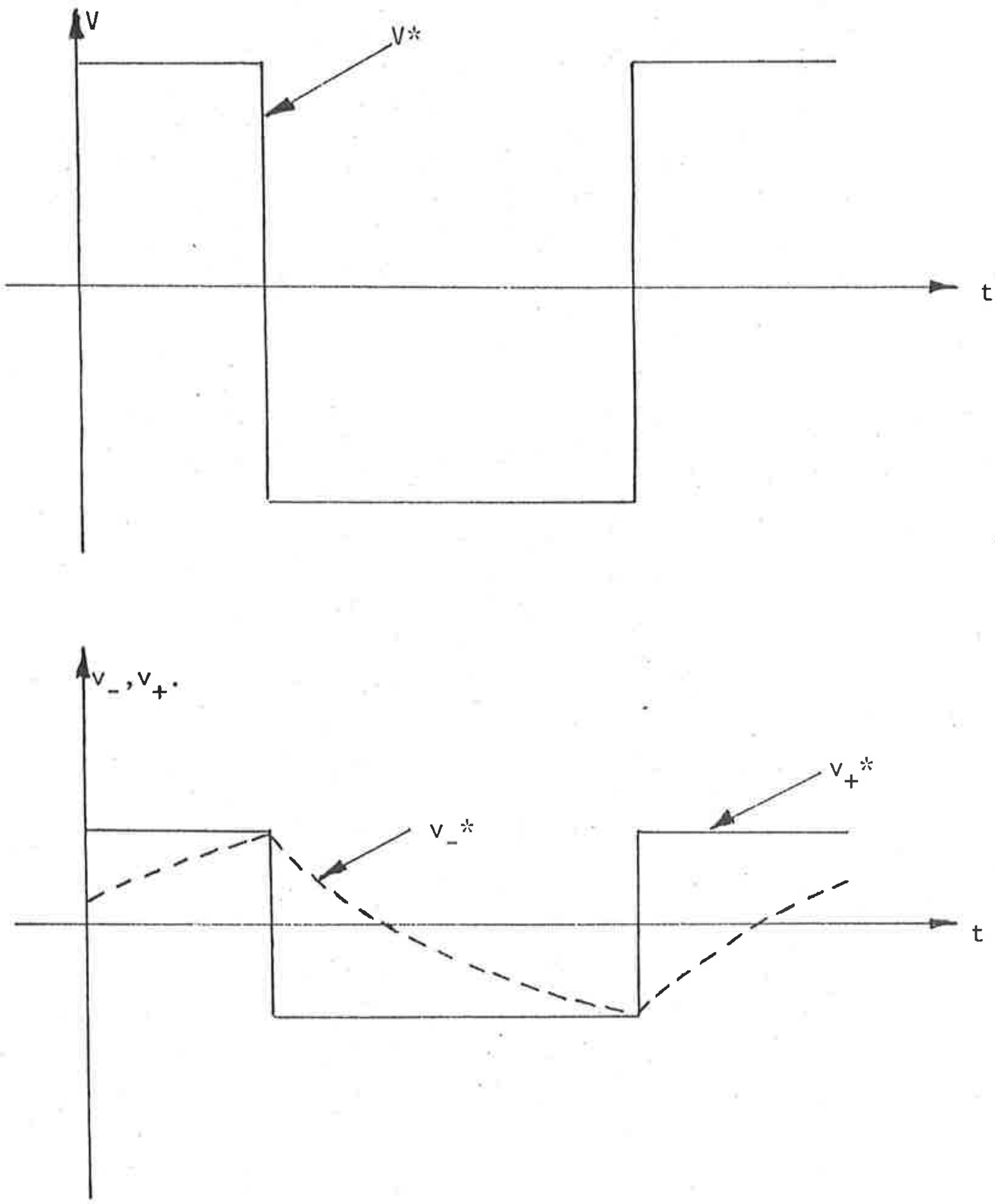


DIAGRAM 6.2.B. Voltage Waveforms for the Circuit of 6.2.A.

*See 6.2.A for the definitions of these voltages.

and (3) The voltage which v- would reach in the steady state.

The potential divider on the noninverting input can be replaced by a precision voltage controlled inverter as shown in diagram 6.3.C. The frequency of oscillation is then determined by a control voltage.

It remains to obtain a logarithmic relationship between this control voltage and the frequency of oscillation.

6.2.3 Obtaining the logarithmic control voltage to frequency characteristic.

The intrinsic relationship between control voltage and frequency is given by,

$$1/f = t = 2T \ln \frac{V_{ss} + v_c}{V_{ss} - v_c}$$

where f is the frequency of oscillation

t is the period of the oscillation

T is the timeconstant of the RC lag circuit

V_{ss} is the steady state voltage to which the capacitor would charge

and v_c is the control voltage.

This equation applies where there is one lag circuit connected to the inverting input. The relationship can be adjusted using two parameters, T and V_{ss}. The relationship can be expressed with the control voltage, v_c, as the independent variable.

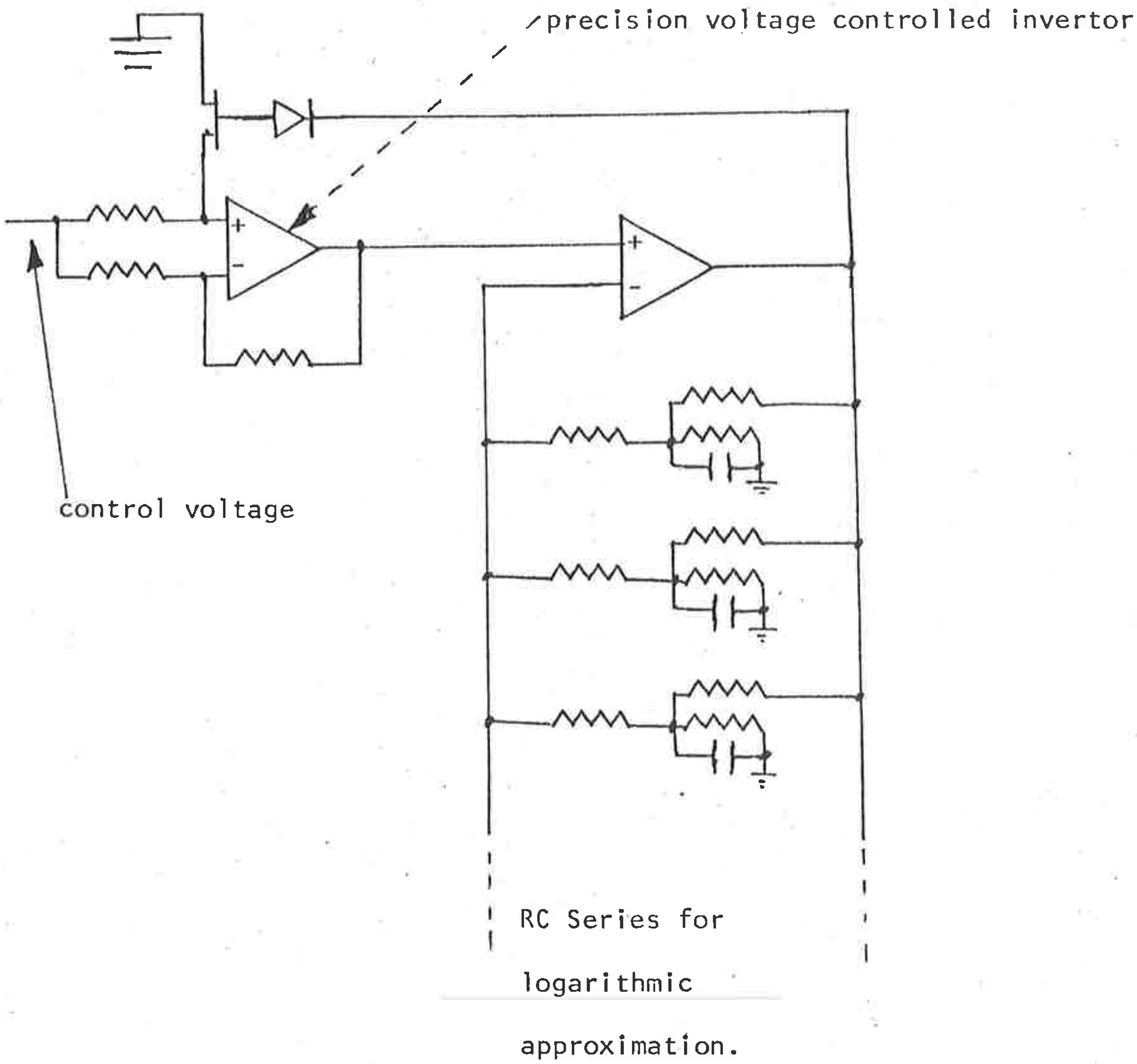


DIAGRAM 6.2.C. Voltage Controlled Circuit (Simplified).

$$v_c = V_{ss} \frac{e^{t/2T} - 1}{e^{t/2T} + 1}$$

A number of first order lag circuits can be combined to give a good approximation to a logarithmic relationship as shown in diagram 6.2.D. In this case

$$v_c = \sum_i V_{ss_i} \frac{e^{t/2T_i} - 1}{e^{t/2T_i} + 1}$$

It is desired that $v_c \simeq k_1 \ln t + k_2$ where k_1 and k_2 are arbitrary constants. k_1 is arbitrary because it is of little consequence what particular voltage increment increases the frequency by a given ratio (within practical limits), and k_2 is arbitrary because it is possible to have a constant voltage offset on the control voltage and still obtain the desired operating characteristic by divider action.

Many approaches to the problem of optimizing T_i and v_{ss_i} to give the desired logarithmic control voltage to frequency relationship were tried. The first approach was analytic, and the second used a computer program to find v_{ss_i} given assumed T_i. Both approaches failed. The analytic solution is far too complex; it may not be possible. The computer program gave no insight into the physical significance of the solution. The successful approach was to use the computer to obtain curves which could be optimized with the aid of visual inspection. In this way the effect of each of the parameters could be visualized.

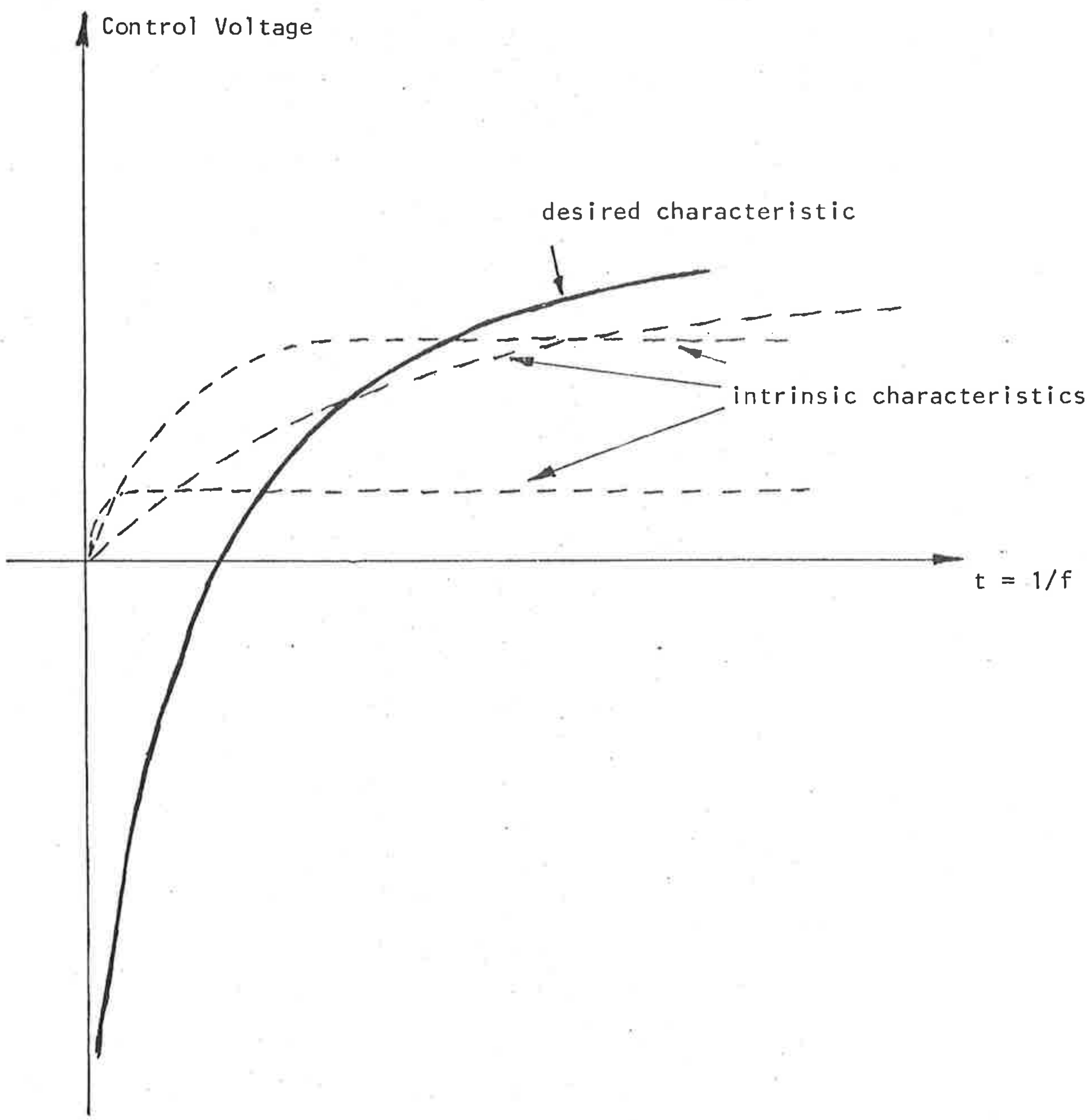


DIAGRAM 6.2.D Control Voltage to Frequency Relationships.

f is the frequency of the oscillation, Hz.

In order to determine the optimum values of v_{ss_i} and T_i it is of advantage to transform the function so that the transformed function corresponding to the desired logarithmic vontrol voltage to frequency relationship is a constant. The substitution u = ln t gives

$$k_3 u + k_2 \approx \sum v_{ss_i} \frac{e^{(e^u/2T_i)} - 1}{e^{(e^u/2T_i)} + 1} = f(u)$$

where k_3 is a constant and directly related to k_1.

Differentiating this function with respect to u gives

$$k_3 \approx \sum_i \frac{v_{ss_i} \; e^u \; \frac{1}{T_i} \; e^{(e^u/2T_i)}}{\left(e^{(e^u/2T_i)} + 1\right)^2}$$

Diagram 6.2.E. gives the form of the curves of f(u) and f'(u).

The advantages of this substitution and differentiation are:

(1) The curve to be approximated is a constant

(2) The approximation functions are relatively simple to visualize

and (3) The value of k_2 (the offset on the control voltage) which is arbitrary because it can be compensated by design does not appear after differentiation.

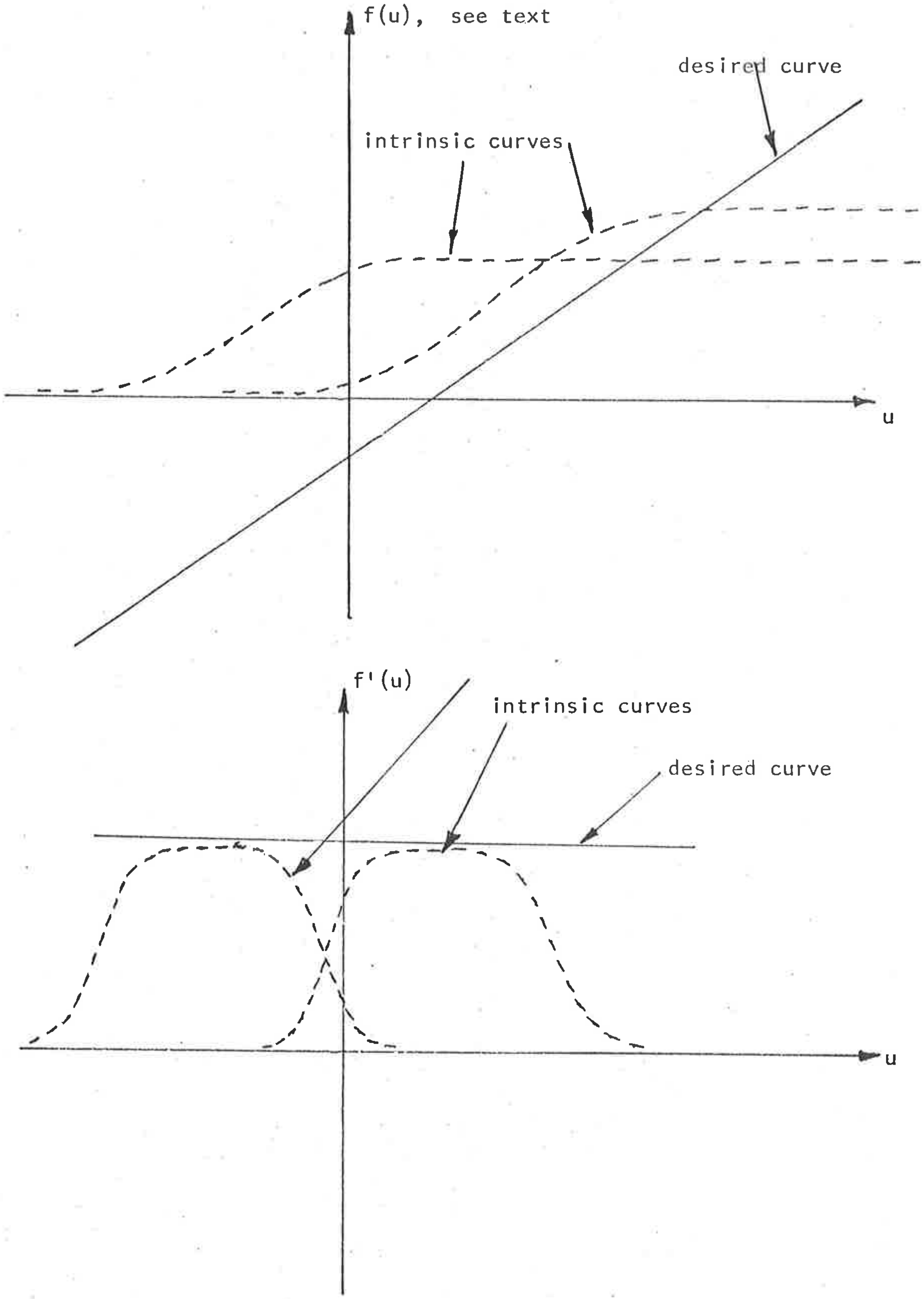


DIAGRAM 6.2.E. Shape of the curves described in the Text.

6.2.4 Calculation of the optimum ratio of timeconstants.

The equation for f'(u) is similar to the equation for $v'_{av}(u)$ which arose in the case of the Sound to Voltage Unit. The differences between the two equations arise because the sound to voltage unit has an output which is the average of the waveform, whereas the Analog Reference Oscillator has a frequency determined by the maximum voltage which determines the end of each cycle.
However the shapes of the curves are similar, and the same technique of solution as applied to the Sound to Voltage unit was used to find the optimum ratio of RC section's timeconstants. The computer evaluated the maximum error as a function of the ratios of the successive timeconstants. Again the optimum ratio was 4, although multiples of this number could be used to improve the accuracy of the oscillator.

6.2.5 Summary of the Analog Reference Oscillator Characteristics

The oscillator uses no nonlinear elements apart from the comparator and this greatly increases the intrinsic stability. The frequency of oscillation never has more than unity sensitivy to the component values. The circuit is intrinsically insensitive to ccommon mode supply voltage variations, assuming that the control voltage varies as does the supply rails. A large frequency ratio is possible because the corresponding control voltage ratio is much less. This means the input offsets of the comparator are much less restrictive. The principle can be used with other applications such as wide band audio oscillators and frequency meters.
(This oscillator was the subject of the 1975 Institution of Engineers Australia Parson's memorial prize lecture.)

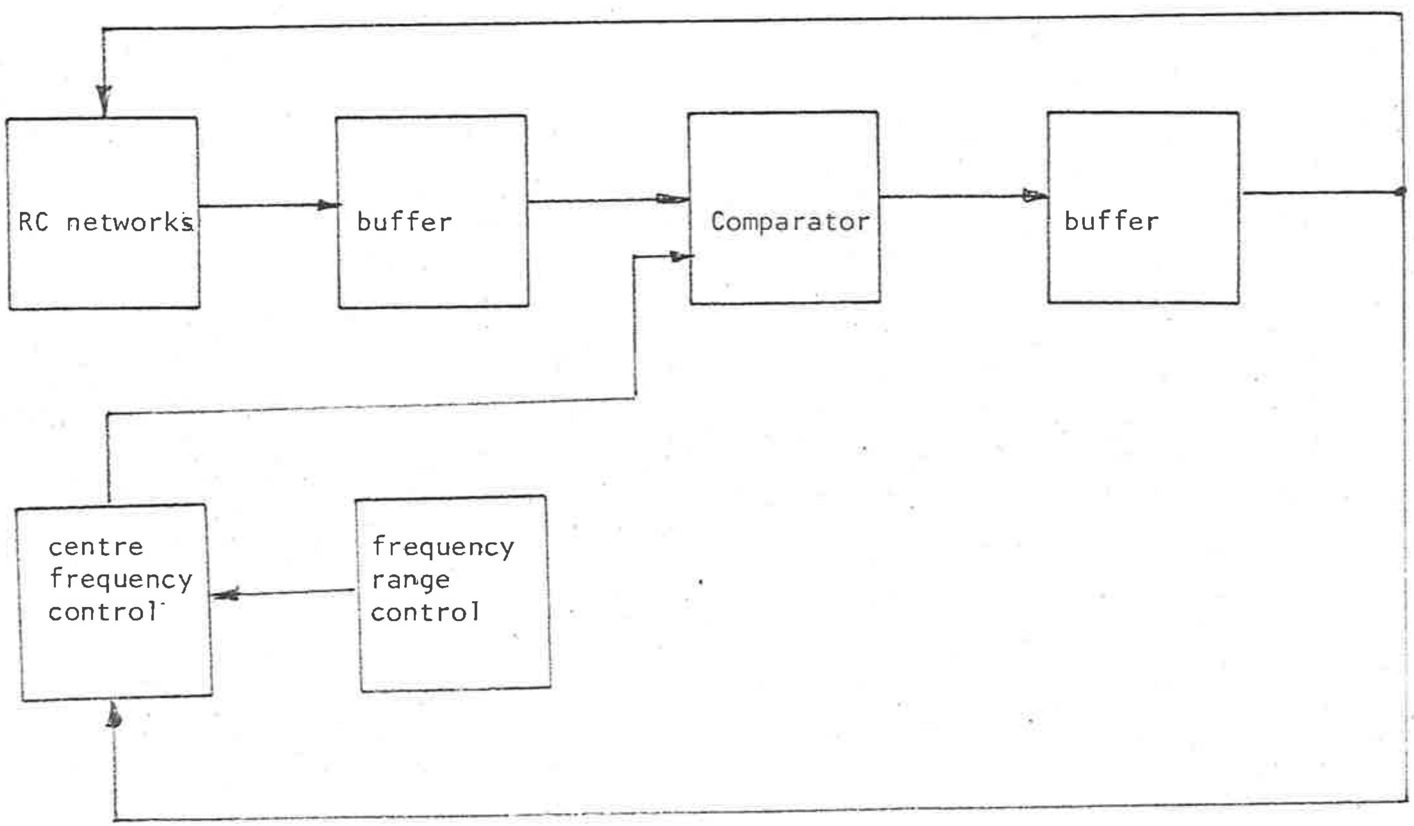


DIAGRAM 6.2.A. Block Diagram of the Analog Reference Oscillator.

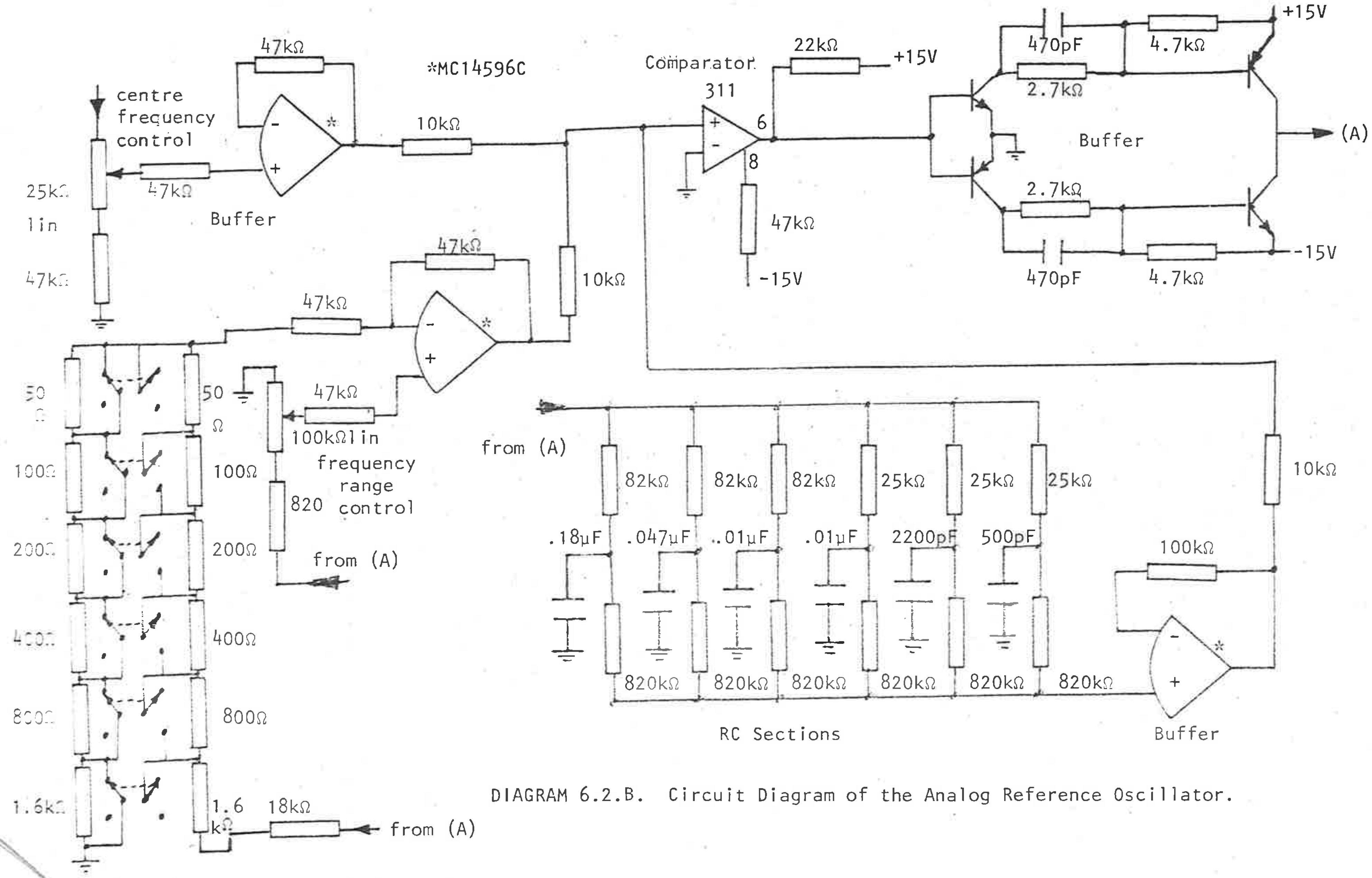


DIAGRAM 6.2.B. Circuit Diagram of the Analog Reference Oscillator.

6.3 The Hybrid Frequency Standard

The Hybrid Frequency Standard combines some of the advantages of analog systems and of digital systems for the generation of a reference frequency. The analog system is often considered more convenient to use. In many applications the continuous control of the generated frequency is considered better ergonomically than a sequential digital setting. However the accuracy of analog devices is less than that available using digital techniques. In electrical systems the analog device used to control frequency is either resistive, capacitive, of inductive. Resistive elements, potentiometers, have .1 % linearity using readily available precision types. The absolute accuracy of a frequency controlled by such a potentiometer may be better than .1 % if the analog control is over a limited frequency range.

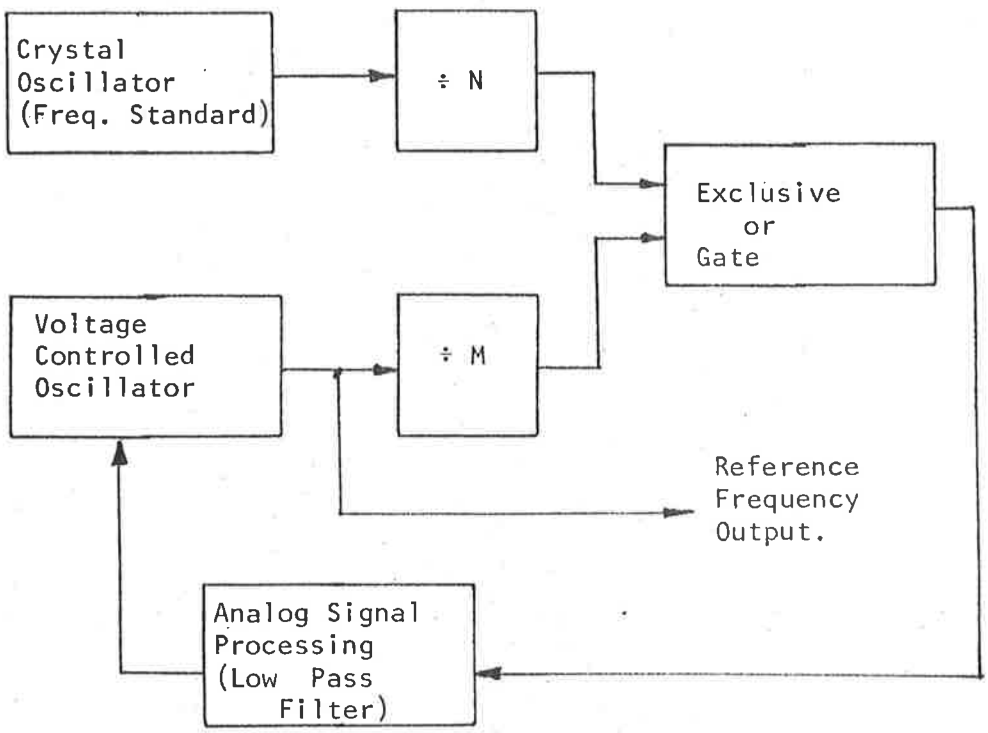
A considerable advantage of a digital system is the use of a standard oscillator of fixed frequency. The desired frequency is obtained using frequency division or phase locked loop techniques. Quartz crystals can be used to provide the standard frequency, and these provide accuracies in excess of .001 % without sophisticated techniques.

The Hybrid Frequency Oscillator enables the use of a quartz crystal standard oscillator, but the dependent frequency has a contimuous relationship to this standard frequency. Analog control of this frequency ratio is achieved using a potentiometer. The circuit is designed so that neither the absolute resistance of the potentiometer, nor the voltage across it alter the reference frequency which is generated. The only time dependent element which determines the frequency is the crystal of the reference oscillator.

A basic phase locked loop system is shown in diagram 6.3.A. The standard crystal frequency is divided by N and the voltage controlled oscillator is divided by M. These signals are then compared in an exclusive or gate. The average output of this exclusive or gate depends upon the phase relationship beteeen the two input signals. This output is filtered and controls the frequency of a voltage controlled oscillator. When locked the frequency of the output exactly equals the reference frequency when it is multiplied by M and divided by N. Both M and N must be integers.

An alternative system to the phase locked loop is shown in diagram 6.3.B, and this is the basis of the Hybrid Frequency Standard. The crystal oscillator and the following digital division network provide a reference timing interval. The impulse generator provides a short pulse at the end of each timing interval which is used to reset the divide by N down counter to N, and to clock the JK flip flop. When clocked the JK flip flop is set to 1 if the down counter has reached zero, otherwise it is set to 0. When the divide by N counter reaches zero the count interrupt ensures this value is held until the next reset pulse to the counter.

The output of the JK flip flop is connected via two analog switches to an integrator. A logic 1 on the output of the JK ensures a positive increment in the output of the integrator. and a logic 0 ensures an equal negative increment. In the steady state 0 and 1 pulses will alternate so that the average output of the integrator is zero. This means that the number of pulses of the output of the voltage controlled oscillator duting two timing intervals is N (corresponding to the JK output of 1) plus N + 1, (corresponding to the JK output of 0). Hence the average number of cycles from the voltage controlled oscillator during each timing interval is N - .5 .



Referency Frequency Output = Crystal Standard Freqnency X M ÷ N

DIAGRAM 6.3.A A phase locked loop frequency synthesizer.

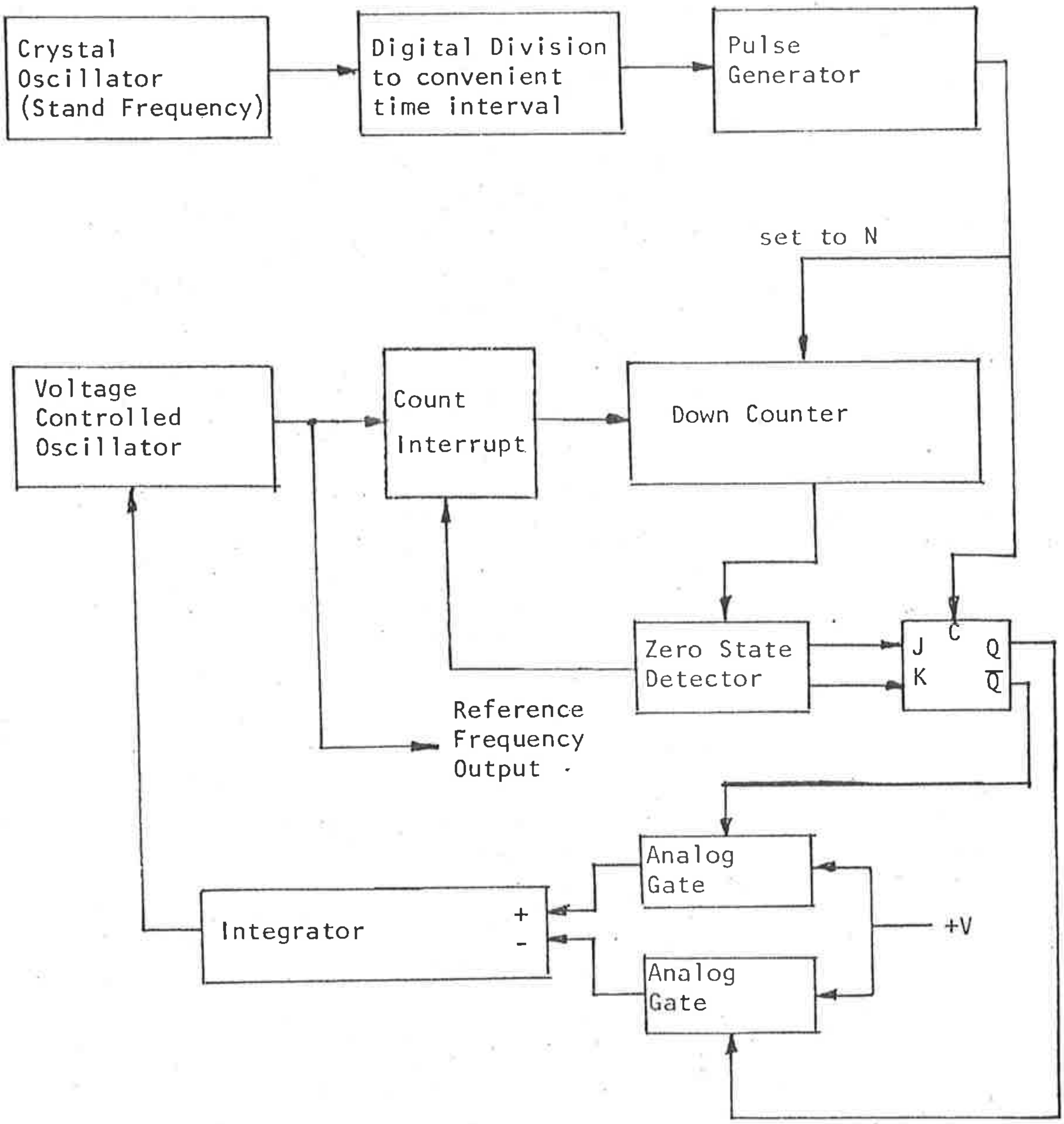


DIAGRAM 6.3.B Alternative system to the phase locked loop for the generation of a dependent frequency from a standard frequency, which is used in the Hybrid Frequency standard.

It is significant that this division is not by an integer, as was the case with the phaselocked loop system.

Each pulse from the analog gate and the JK flip flop is integrated and therefore changes the frequency of the voltage controlled oscillator. This means the generated reference frequency is continually varying about a mean value. The amount of this frequency deviation can be made arbitrarily small by decreasing the integration rate. A more sophisticated low pass filter could be used to minimize this frequency variation if the application of the oscillator restricted the time available for a simple integrator to adjust to a desired frequency.

The analog control of the Hybrid Frequency Standard is introduced by providing an offset of the integrator. This control determines the ratio of 1's from the JK flip flop to the ratio of 0's in the steady state. Diagram 6.3.C gives the block diagram of the Hybrid Frequency Standard.

If the potentiometer is set of 50% then the number of 1's from the JK flip flop must equal the number of 0's as was the case with the case with the system in diagram 6.3.B. However if the potentiometer setting is greater than 50%, the number of 1's from the JK must exceed the number of 0's from it in order to ensure the average output from the integrator maintains a constant voltage level.
The potentiometer setting can control the ratio of 1's to 0's from 100% (zero count reached every time), to 0% (zero count never reached). This gives analog control over the least significant bit. The adjustment to exactly 100% or 0% is liable to instability because any frequency overshoot exceeds the range of control. This effect can be overcome by removing the zero state detector from the least significant bit.

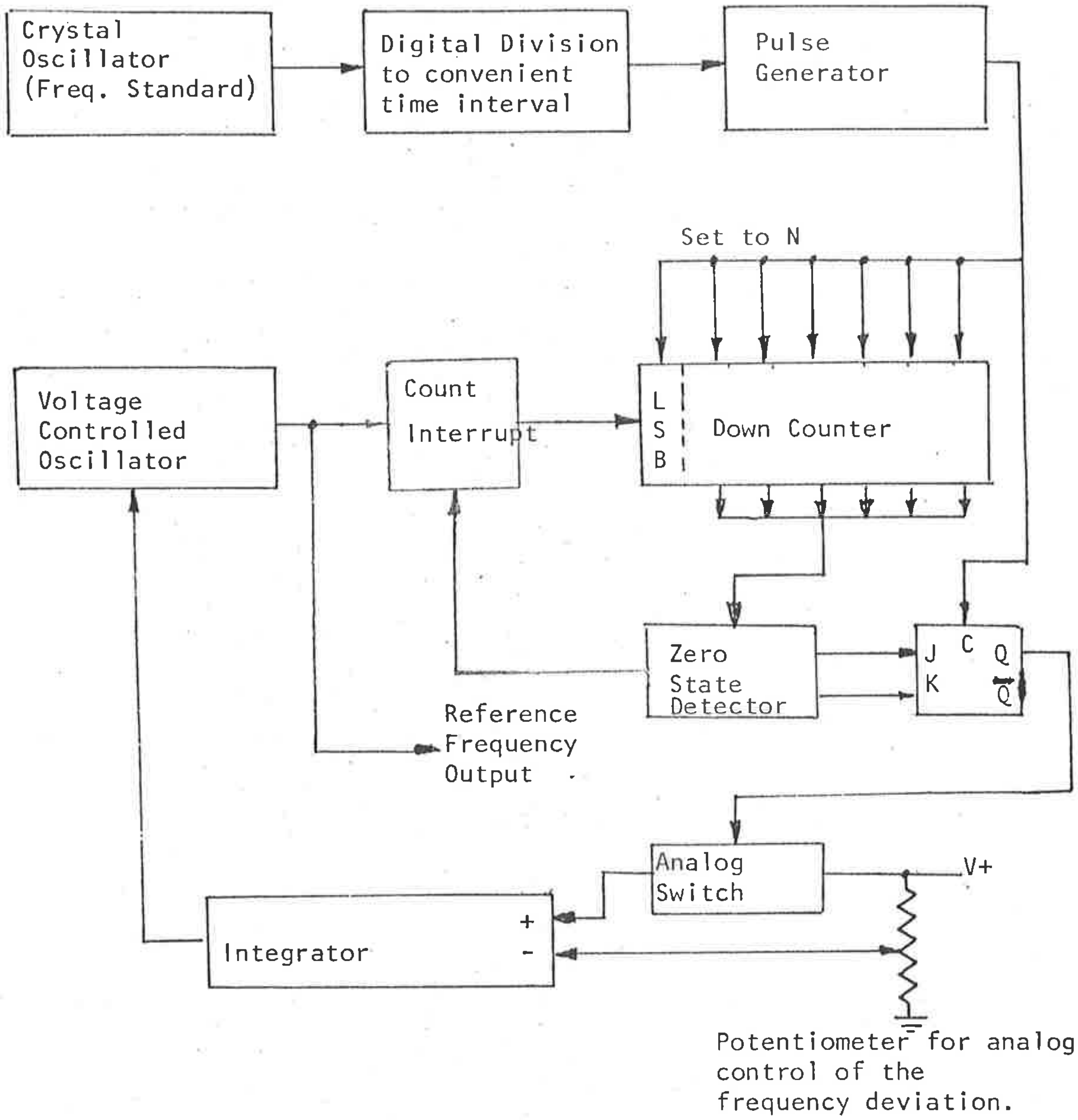


DIAGRAM 6.3.C Block Diagram of the Hybrid Frequency Standard.

This means the analog control voltage range is now over the second least significant bit. The least significant bit is still set to N when the divide down counter is reset. Now that the analog control is over the second least significant bit, overlap of the analog control ranges is provided. This means continuous control of frequency is possible.

The principle of the Hybrid Frequency Standard was realized using an EAI 180 analog/digital computer. The computer schematic is given in diagram 6.3.D.

There are numerous features of the principle. The crystal is the only energy storage element which determines the frequency of the oscillator; the integration rate merely alters the rate of attainment of the frequency. The analog control is provided by a potentiometer and the circuit is such that the linearity of the potentiometer is the only important property; its absolute resistance does not change the frequency of the oscillator. The integrator can be arranged so as to not load the potentiometer which would otherwise alter the lineraity of the potentiometer setting. Diagram 6.3.E gives a suitable integrator circuit. The same voltage is applied to the potentiometer as is applied to the analog gates. This means the voltage level does not alter the frequency from the oscillator; only the rate of adjustment of frequency is altered.

There are three additional sources of error with the oscillator which must be kept within the required tolerances by suitable design. The accuracy of the system will be degraded if the input currents to the operational amplifier are significant compared to the operating current levels. In this design the input currents are 25 nA

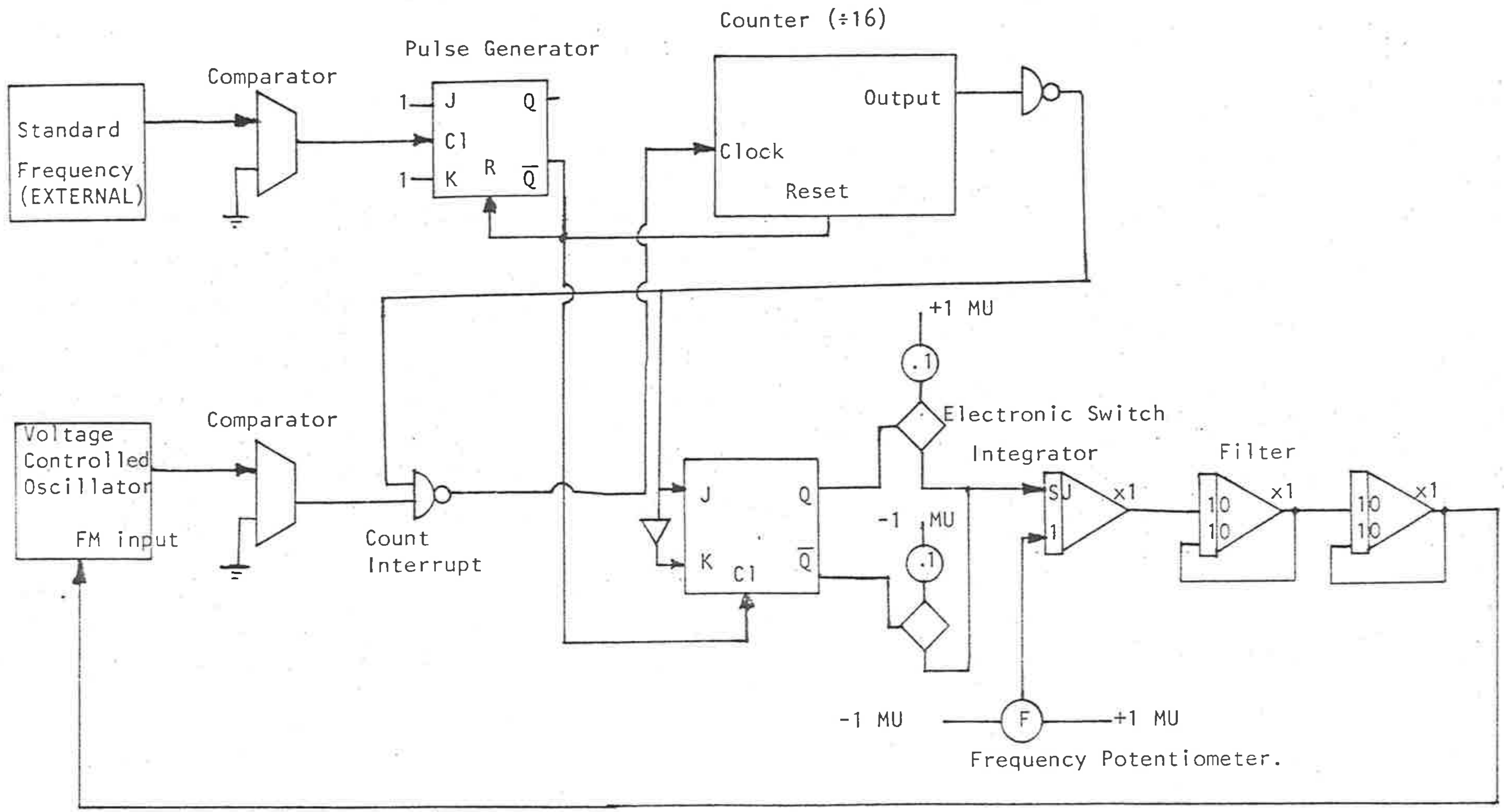


DIAGRAM 6.3.D. Analog computer realization of the Hybrid Frequency Standard.

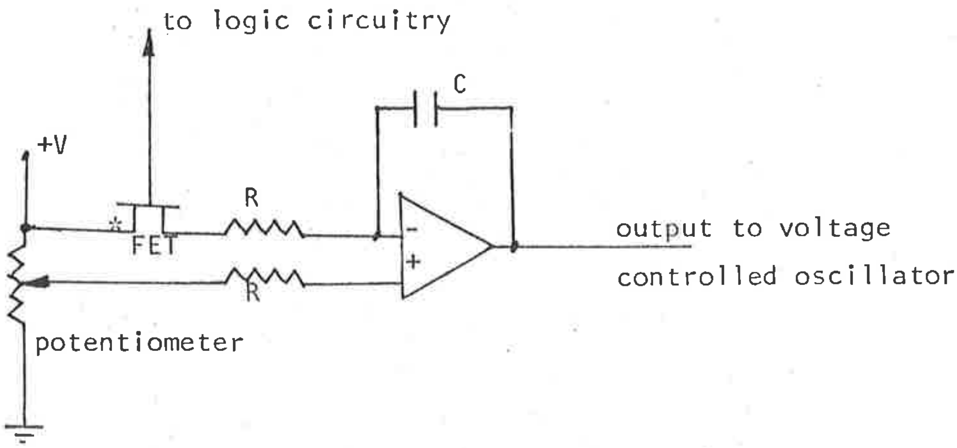


DIAGRAM 6.3.E Differential integration circuit which does not load the precision potentiometer, and uses only one voltage polarity, +V.

*In this circuit a P channel FET would be required. The source is marked (*), and the gate voltage is then with respect to +V. It follows that the gate is never forward biased so no significant offset current flows into the integrator through the FET.

for the μA 777 precision operational amplifier, and the operating currents are 50 μA. Also the duration of the reset pulse must be small compared to the count rate of the voltage controlled oscillator. If the reset pulse is present during the count from the voltage controlled oscillator the count will be missed and a frequency error will result. In this design the pulse duration is 100 ns and the count rate from the voltage controlled oscillator is 100 μs. The switching time of the analog switch must be small compared with the duration it is switched either on or off. The average current through the switch must be proportional to the average time it is switched on. Typical switching times of 10 ns are available, and a switching duration of 100 ms is suitable for this application.

Further applications for the principle of the Hybrid Frequency Standard include a radio frequency oscillator where the stability comparable to that of a crystal oscillator is required but analog control over a limited frequency range is preferred. The principle was developed for the audio frequency application, and is believed to be novel.

NOTES FOR CHAPTER 6

1. The musician's requireement for even tempered scales is discussed in reference 21, pp 82.

CHAPTER 7 CONCLUSION

The layout of the instrument is given in Diagram 7.A. Each unit was tested individually and the cabinet for the units was completed.

There have been many recent improvements in electronic engineering components which have direct application to most of the designs of the synthesizer. Integrated signal multipliers were already available when the Filter was designed though not when the signal multiplier was built. Inexpensive digital and analog interfaces and storage are now readily available, suggesting digital filtering as an alternative approach. High quality acoustic delay can now be provided, and soon charge control devices will simplify these designs even further. The harmonic generator could now be made using 20 operational amplifiers at a total basic component cost of $5. At the time the transistors were chosen each operational amplifier was worth twice that price. Inexpensive digital displays have made the provision of a standard reference frequency outmoded in this application.

The microprocessor has direct application in the storage and generation of sounds, It would be useful to provide the co-ordinating role of the switching matrices. Three microprocessor Arithmetic Logic Unit integrated circuits are now available for the origional cost of the switches, however it must be remembered that peripherals would be required. Possibly the greatest scope for revolutionary design of a music synthesizer using a microprocessor is the programmed instrument which is described in chapter 1. This relaxes the processor time requirement but demands a comprehensive display of the sound for the musician. This display must be easily interpreted, changed and recorded.

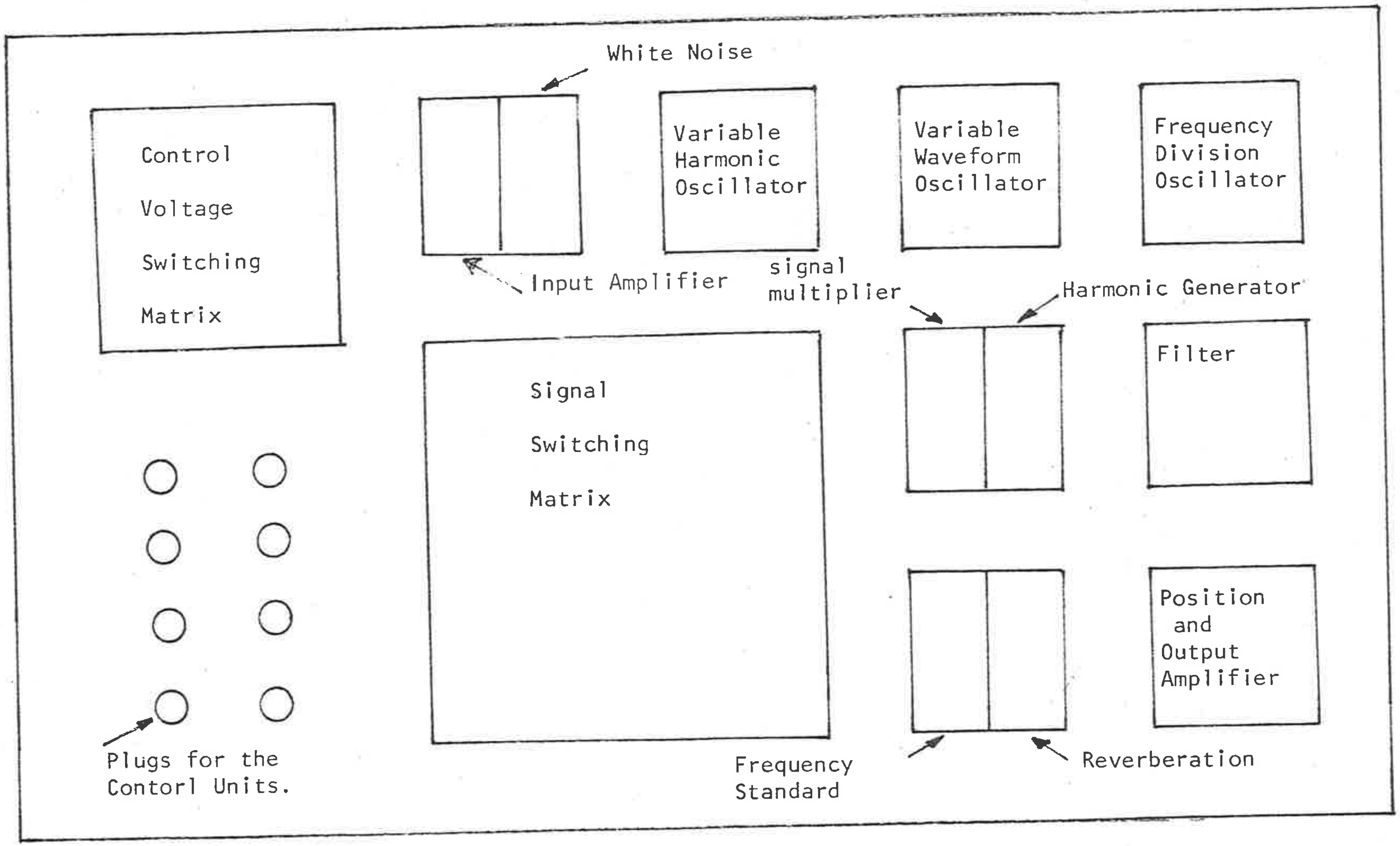


DIAGRAM 7 A. Layout of the Complete Instrument.

These technical advances are comparable with those which provided the first integrated circuits, and ofcourse there will continue to be further development. The fact which will endure is that if a musician wishes to select one of a number of sounds, then he must select one of that number of physical movements. The design requirements of a synthesizer must take account of this fact and its implications. These implications are discussed in Chapter 1, and include the predictability of the sound sources and sound manipulators as well as the use of many degrees of freedom by the control unit. It is tempting to provide a complex system: A large computer with a special language which makes any sound available to the musician. The necessity of selecting a sound using a teletype, slowly punching in mathematical equations, makes the selection of a particularly subtle sound a physical impossibility. It is the interface with the musician which can unduly limit this type of musical instrument. The challenge will always be to design a synthesizer which will take full account of this consideration.

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APPENDIX. DYNACON PRESSURE SENSING MATERIAL.

The manufacturer's description of the Dynacon pressure sensing material is given. This material was used to sense pressure applied to the Binary Keyboard (Section 5.3). It was found by measurements on the three types of material; Dynacon A, B, and C, that Dynacon C, had the most suitable properties.

*DYNACON MATERIALS KIT

117 Fort Lee Road, Leonia, N. J. 07605 Tel: (201) 947-0106

Enclosed in this KIT are three samples of Dynacon.
Sample A is conductive lineally, and may be used as an electrode for the other pieces, and as electromagnetic shielding.
Sample B is a quick-switch material that can be used for simple on-off pressure devices such as alarms, sensors.
Sample C is the new pressure-conductive polymer which can sense weights and pressure and control response switching.

Description of the Material and its Properties

Material Properties:

The material behavior can be described as follows:

1) When at rest, the conduction of electric current between any two points is zero. (10 Megohm impedance.) This is true for voltage levels under 40V, above which the material will start to break down.

2) As a stress is applied, the conductivity of the affected area is proportionally increased in the direction of the stress. The stress may be compressive or tensile, and the polarity of the applied voltage does not aggect the magnitude of the current through the material. As the pressure is released, the initial electrical characteristics of the material are restored (no conduction).

3) In addition to Dynacon's response to mechanical stresses, there are other parameters that affect its electrical characteristics, among which are the following:

a) Thermal response—The material has a negative dI/dT coefficient when under pressure.

b) When the material is kept under a constant stress, the voltage applied will affect conductivity in a positive irreversible manner. For example, if the material is carrying 0.100 amps at 5.0 V. and the voltage is increased to 10.0 V. and then decreased again to 5.0 V., the current will now be somewhat higher than 0.100 amps. Thus the material exhibits retentive (or memory) characteristics. When the stress is relieved, the initial set of conditions is restored.

c) If the material is stressed from rest to a given pressure, so that, for example, it conducts .1 Amp at 10 V., and then it is exposed to an electromagnetic field (not necessarily large) the current will be increased. As the field is removed, the current will remain at the last value. Several levels of conductivity can be achieved by this process, and the last one will be the governing one. Thus, again the material exhibits retentive capability. The initial set of conditions can be restored by removing and reapplying the stress.

d) Time effect. If the material is simply stressed between two plates, it will slowly deform with time, and the electrical characteristics will change as well. Normal setting times are in the range of 15-60 seconds.

APPLICATIONS — 1

Applications requiring an **insensitive switch** action:

a) Door mats

b) Intrusion alarms

c) Limit switch

The sensing element can be made by sandwiching a piece of Dynacon B between two rigid conductive surfaces (copper clad, copper plates, etc.) and using it as a normal switch for voltages below 40 V and currents below 1 Amp. Fig. 1 shows a typical application

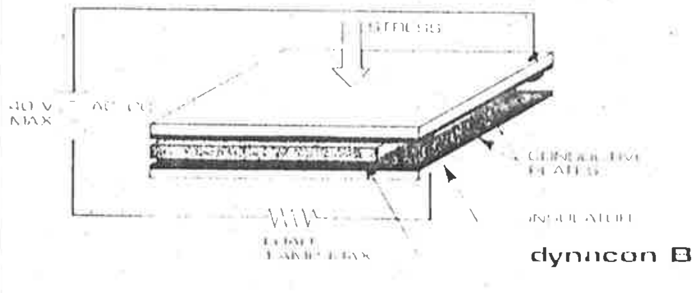


FIG 1

APPLICATIONS — 2

Sensors requiring a **sensitive switching** action:

a) Shock sensors
b) Weight switches
c) Pressure switches
d) Intrusion alarms
e) Limit switches

Sensors of this type can be made by sandwiching a segment of Dynacon B between two rigid insulators, but the contacts are in the form of a grid. Thus the affected area is extremely small and the resulting switch very sensitive. See Fig. 2 for an illustration.

For further refinements and increased sensitivity a "needle" can be used (low pressures only) as illustrated in Fig. 3.

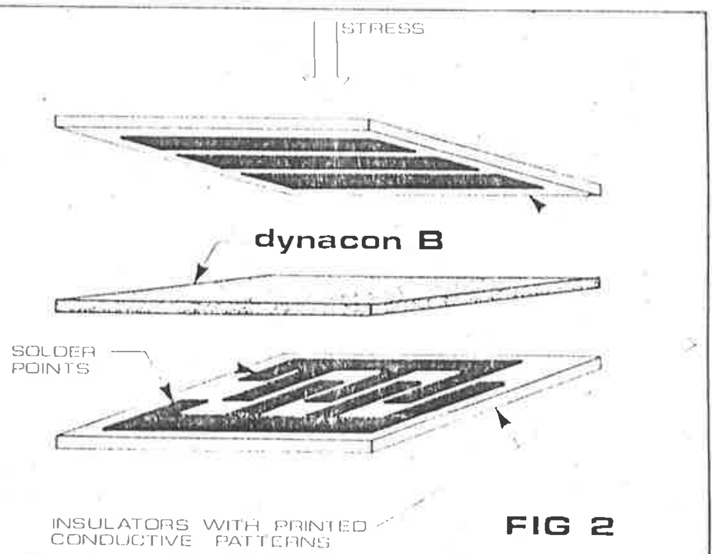


FIG 2

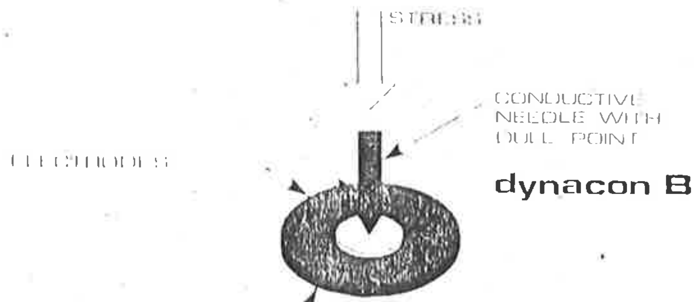


FIG 3

APPLICATIONS — 3

Linear pressure sensors:

Linear pressure sensors can be achieved by replacing Dynacon B with Dynacon C in the two previous cases. The replacement will give a linear sensor in case 1 and a very sensitive linear sensor in Case 2. When used in this manner, good repeatability and settling times cannot be obtained, but for applications with constant feedback or intermittent use this concept works extremely well. Some applications include:

a) Manually operated motor speed controllers.
   1) Handdrills
   2) Saws
   3) Sewing machines

(See Fig. 4 for illustrations)

b) Pressure sensor for press, e.g., Punch press.

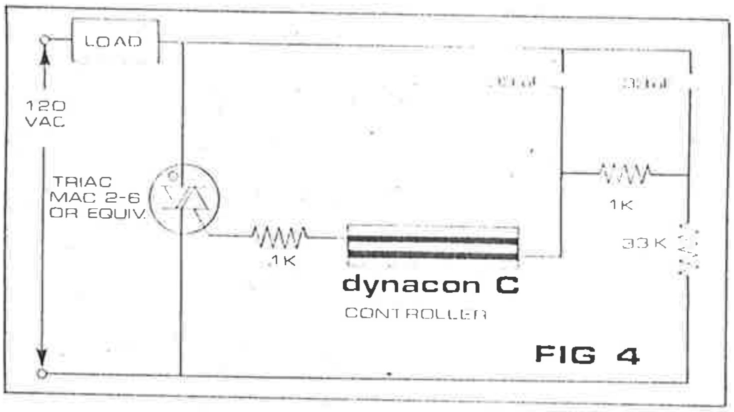


FIG 4

APPLICATIONS — 4

(Fig. 5)

Pressure Transducer Applications:

The repeatability of Dynacon C material can be greatly increased by relieving the bulk of the pressure applied to the material by the use of a material with a larger modulus of elasticity. This can be achieved by several methods. Figure 5 illustrates such a method, however, care must be taken in the actual construction to keep the dimensions of the Dynacon C material and the holder very close. A slight deviation in either direction

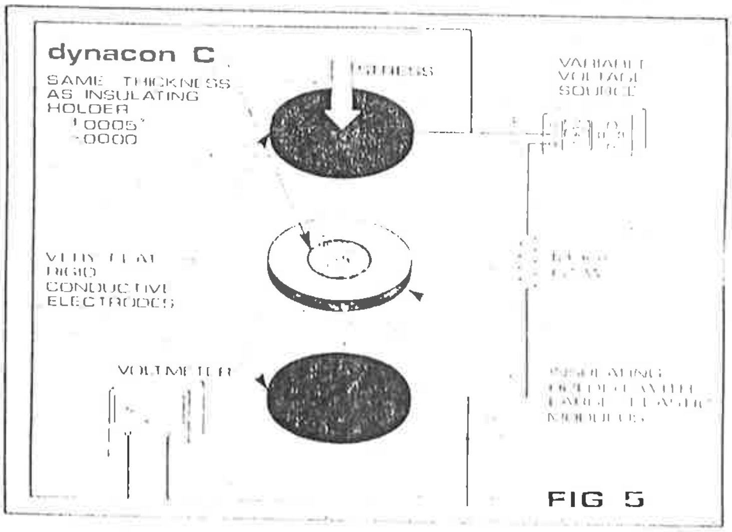


FIG 5

will cause the sensor to generate erroneous readings. Also, when using Dynacon C in this way care must be taken to provide a constant voltage supply and suitable R. F. Shielding. The range and sensitivity can be adjusted by changing the dimensions of the holder and/or its modules of elasticity.

The enclosed sample of Dynacon C has a fairly large hysteresis, and is not suitable for precise measurement. It is excellent for determining pressures, especially in situations where the environment is hostile to delicate electrical equipment, or in inaccessible positions as between rollers, or under the acid mud of a pipe line. We presently have under test a much more precise material, DYNACON D. See graph for loading data.

APPLICATIONS — 5A

(Fig. 6)

Memory Applications

With the Dynacon C material at a given pressure and voltage, record the current through the circuit. Then increase the voltage by at least 50% and reduce it to its original setting. Do not change the applied pressure, and note the new reading. Note that this is higher than the original current. Now release the pressure and reapply it. The current will be the same as in the original case. This process can be repeated as many times as desired, it is non-destructive.

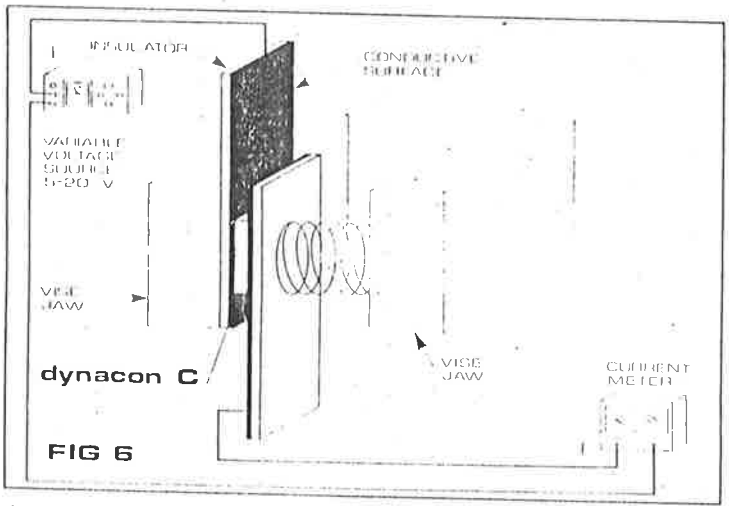


FIG 6

APPLICATIONS — 5B

Memory Applications

In APPLICATION 5A the current was a function of the voltage applied to the Dynacon C material. In this case, the current will change by the application of an electromagnetic field.

Start by applying a constant voltage and pressure to the Dynacon C memory as in Fig. 7. Record the current. Now close the switch one or more times. The current will change (increase) when the switch is released, because of the fields generated by the coil from the transient condition. The original condition can be restored by relieving and reapplying the stress.

NOTE: Oscillating the stress without fully removing it will also restore the original conditions.

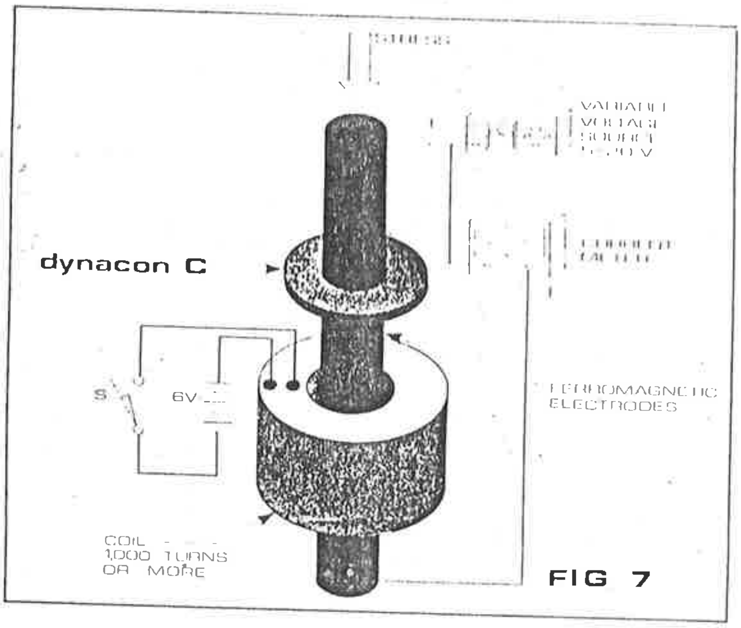


FIG 7

WEIGHT SCALE APPLICATION

Description of the machine:

a) SENSOR:

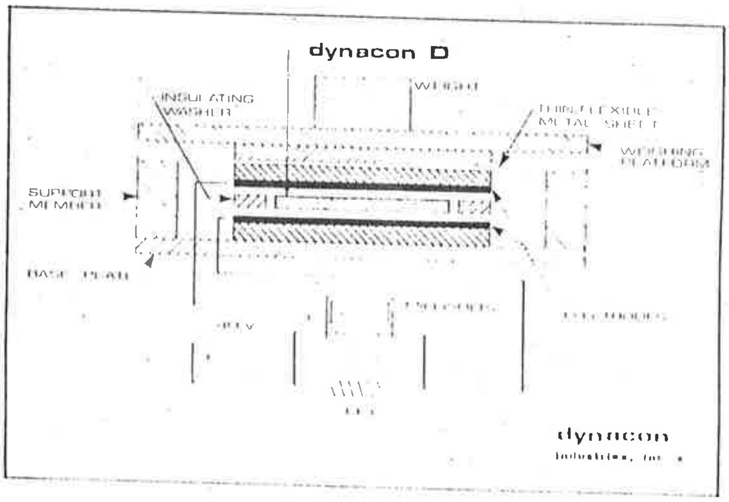
The sensing element consists of a segment of Dynacon D material encapsulated in a nonconductive elastic holder of the same thickness. In this fashion, the bulk of the pressure is absorbed by the holder, while the Dynacon segment senses dimensional changes.

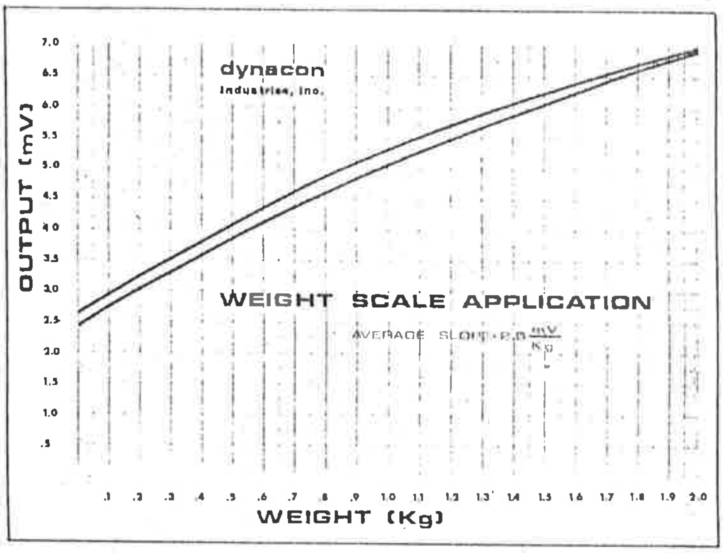
b) MECHANICAL:

Pressure is applied to the sensing element by two electrodes, one which is fixed within the frame of the machine, and the other is floating. The floating element is attached to a thin flexible sheet of metal so that it moves in a parallel manner. The pressure is directly transferred from the top of the sheet to the electrode.

c) ELECTRICAL:

The Dynacon D sensing element is connected in series with a resistor and power source. Current measurements are indirectly obtained from the voltage drop across the series resistor. A digital voltmeter was used in this experiment.





Please note that the material used in this application is Dynacon D, which is not yet in production, and has not been included in the kit. Dynacon C can be used, but results are not as precise. Samples of Dynacon D will be made available at a later date.

The graph of Weight Scale Application refers to measurements made with Dynacon D.