TOPICS ON HUMAN LOCOMOTION.

1. MULTICHANNEL TELEMETRY OF ELECTROMYOGRAPH SIGNALS.
2. PROGRAMMED FUNCTIONAL ELECTRICAL STIMULATION.

BY

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<u>DECLARATION</u>.

I declare that this dissertation is my own unaided work and has not previously been submitted for a degree to any University.

Ari Ziskind

Signed by candidate

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CHAPTER 1.

STATEMENT OF THE PROBLEM.

1.1 INTRODUCTION.

A detailed understanding of the human locomotor system is essential before work can be done to rectify damage to any part of it. Although the integrated locomotor process, which is simply the forward propulsion of a person, seems a very simple activity, the individual parameters are many, and interwoven with a tremendous amount of complexity. Slight variations in the shape of a bone, or the ineffectiveness of a single muscle might completely unbalance the system and make the simple walking process a very difficult or painful affair.

A detailed knowledge of these parameters is necessary for professionals who attempt to assist amputees, paraplegics, deformed or even only rheumatoid people to walk again or just move around. Among these professionals is the surgeon who undertakes surgical procedures to improve the function of damaged sections of the system. There is the physiotherapist who supplies addaptive functional training to the patient; and then the bio-engineer who designs limb prostheses or splints for support and assistance or maybe even a muscle stimulator for a paralysed muscle.

1.2 HUMAN LOCOMOTION.

The parameters studied in human locomotion are many and varied. They include such aspects as the variation of the angles subtended between limb segments, the curves joints trace through space during walking, the forces exerted on the sole of the foot and the phasic activity of the muscles. Some of the monitored parameters might seem extremely isolated and of no specific use, but gait abnormalities might be shown up very markedly by an abnormal variation in certain parameters. This factor could be used to advantage as a clinical tool in the assisted diagnosis of certain locomotor ailments.

1.2.1 PRINCESS ALICE HOSPITAL RESEARCH PROJECT.

The locomotion research project of the Department of Bio-engineering and Medical Physics is at present stationed at Princess Alice Hospital. At the moment the main aspects of this project are:

a) E.M.G. STUDIES.

The system has six available information channels. Four of these are used for e.m.g. information and two for footswitch patterns to reference the e.m.g.s. Facilities exist to get the subject to walk at fixed speeds or fixed pace frequencies. The monitoring is done either on ultra-violet paper or on a fourteen-channel magnetic tape-recorder, of which, however, only eight can be used.

A computer programme has been developed for the analysis of this data and the determination of the phasic activity of the muscles. A laboratory technical report by Hershler, C., McConnell, V.A., and Milner, M. (1972), contains a detailed analysis of this work.

b) STROBE PHOTOGRAPHY.

A stroboscope with a long exposure camera has been set up which traces the trajectories of certain fixed points on a subject through space; viz., shoulder, hip, knee, ankle and toe. A computer programme has also been developed in the Department which uses these input data for determining the variation of angles subtended between the various limb segments. A detailed analysis of this work appears in papers by Milner, M., Wilberforce, C.B.A., and Brennan, P.K., (1972), and Hershler, C., McConnell, V.A., and Milner, M. (1972).

1.3 E.M.G. TELEMETRY.

The system for measuring e.m.g.s being used at Princess Alice Hospital has a differential amplifier mounted next to each muscle, and a set of long leads connected from the patient via a freely swinging beam to the recording apparatus. Section One of this thesis was the design of a telemetry system to bypass these connecting wires. (A full investigation of the advantages of using telemetry here is discussed in Chapter 4.)

1.4 MUSCLE STIMULATION SYSTEM.

It is possible in the case of paralysed patients, where paralysis is due to a nervous dysfunction, to bypass the normal nerve channel and stimulate the muscle directly with an electrical stimulus. A certain number of practical systems have been built for hemiplegic patients, i.e., people who are paralysed on one side only. In the design of a muscle stimulus system, the engineer relies heavily on the type of information mentioned in the previous paragraphs. (A detailed discussion of this work is given in Chapter 8.)

Several systems have been designed for the purpose of stimulating the tibialis anterior muscles of hemiplegics. Section Two of this thesis contains the development of this type of system to satisfy the stimulation requirements of the gastrocnemius and quadriceps muscle groups.

CHAPTER 2.

THE E.M.G.

2.1 INTRODUCTION.

Perhaps one of the most fascinating aspects of modern science is the unity it brings to all knowledge. The fact that a muscle movement may manifest itself as an electric signal reveals links between two branches of science which might outwardly seem to be completely independent. Perhaps this justifies the changing of an old saying to "No science is an island." It is of importance then, for a sufficient understanding of this thesis and the reasons for doing it, to realise exactly how the above link takes place.

2.2 THE CELL.

Basically the human cell consists of a nucleus surrounded by the cytoplasm. The whole cell is then separated off from the surrounding fluid by the cell membrane. Due to the structure of the membrane, a resting potential of approximately 80 mV. exists across it making the inside negative with respect to the outside. The actual mechanisms involved are rather complex and still not fully understood, but an idea of the mechanism might be gained by considering it in the following simple manner.

Both the fluids on the inside and the outside of the cell membrane are rather complex electrolytic solutions containing both Potassium (K+) and Sodium (Na+) ions. Now, there exists two "pumps": a Sodium pump, pumping all the Na+ ions out of the cell and a Potassium pump, pumping all the K+ ions into the cell. The ions then diffuse back in the opposite direction due to the concentration gradients set up by these pumps. The permeability or conductance of the membrane is, however, less for Na+ ions than K+ ions, allowing less of them back in, and so in the equilibrium situation, we have an excess of positive ions on the outside of the membrane creating a potential across it.

2.3 THE ACTION POTENTIAL AND ITS PROPAGATION.

When the membrane is excited or stimulated in some fashion, e.g., chemically or electrically, the permeability of the membrane to Na+ ions increases several thousand-fold and the Na+ ions rush in, because of the resting potential, causing the inside to go positive with respect to the outside. Due to the electrical gradient thus formed and the fact that now the membrane also becomes even more permeable to K+ ions, these rush outwards to bring the potential back to its resting potential. At this moment, then, the membrane loses its increased permeability and after a short time the concentrations on both sides of the membrane return to their initial values. The whole potential change lasts a fraction of a millisecond and varies slightly with temperature.

The effect may be seen clearly in Fig. 2.1.

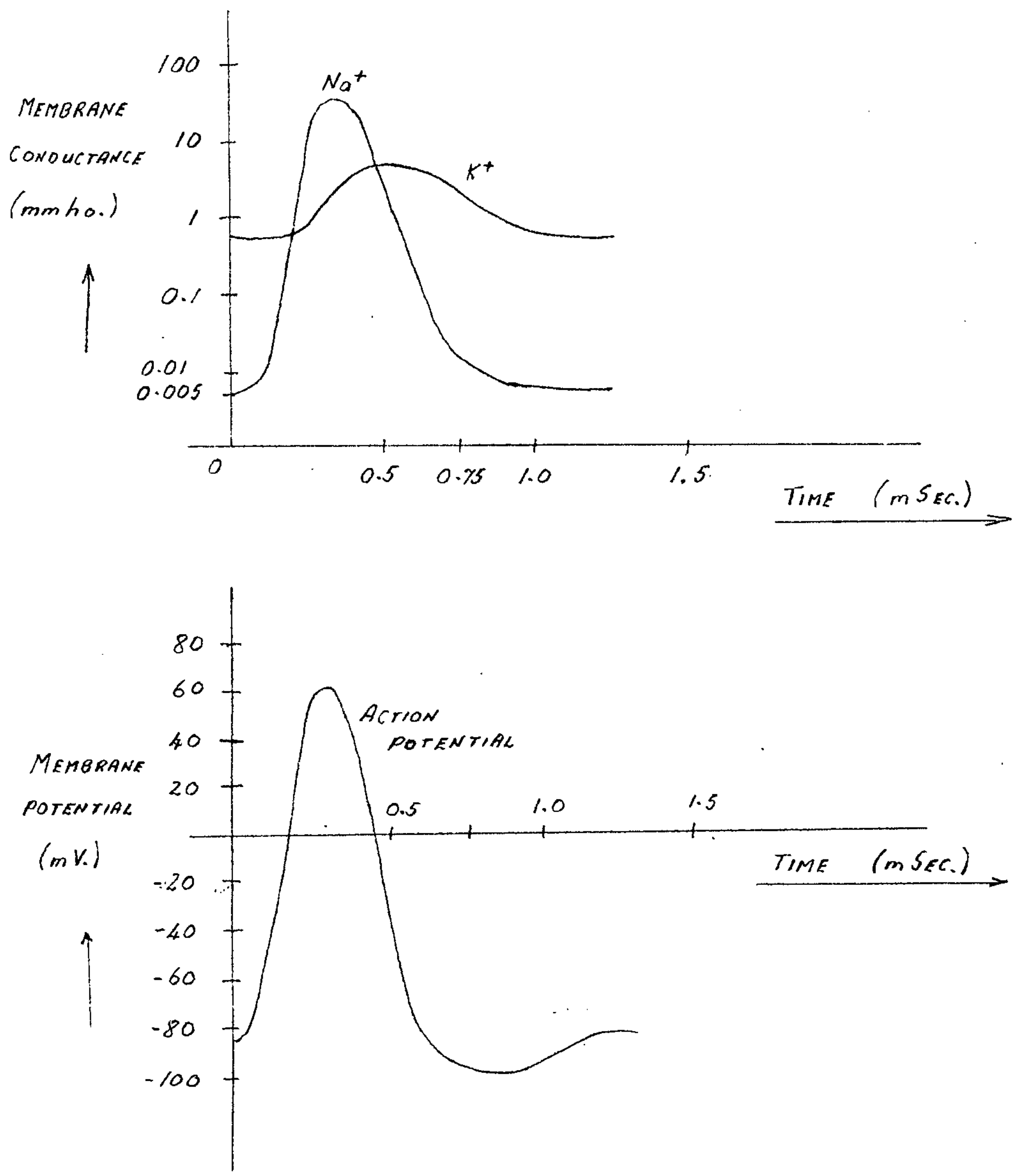


Fig. 2.1

An important property of the action potential is its all-or-none law. That is, it has a certain set magnitude and shape independent of the size of the stimulus, but there is a certain critical level of stimulation before it occurs.

Once an action potential is generated at a certain point on the membrane, it automatically propagates along the membrane in both directions. Referring to Fig. 2.2, we can see that due to the presence of the action potential at a point on the membrane, a local circuit of current flow between the depolarised and the polarised areas of the membrane occurs. This current flow now changes the permeability of the previously unaffected membrane area and so the action potential is propagated along the membrane surface.

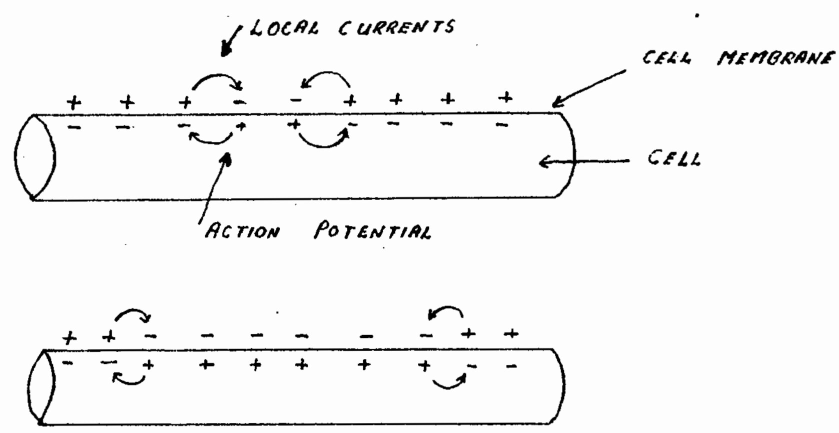


Fig. 2.2

2.4 THE NERVE FIBRE.

The nerve cell, like all other cells, consists of a nucleus and a membrane. However, its shape is quite unique in that it consists of a cell body with a long thin snake-like fibre, called an axon. These axons may be up to three metres in length. Little fibrils or dendrites protrude from the cell head and it on these that the axons of other nerves connect. At the end of this line, the final axon is connected to the muscle fibre which it controls.

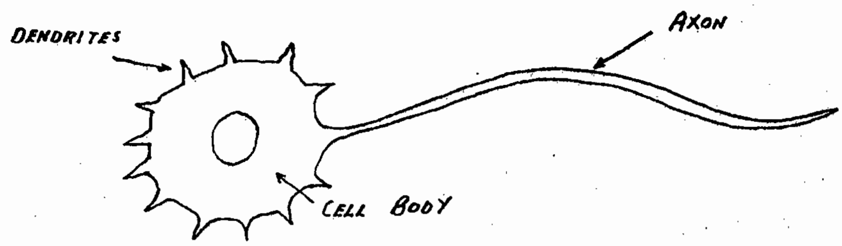


Fig. 2.3

The nerve cell propagates its impulses in the manner described above.

2.5 THE MUSCLE FIBRE.

The muscle fibres are about 30 mm. long with a width of approximately 100 microns. Here again the formation and propagation of the action potential is the same as in that of any other cells, except that its duration is about 5 to 10 msec., compared to 0.5 msec. in the case of the nerve fibre.

Now as the action potential passes along the muscle fibre, it causes it to contract. This contraction occurs about 3 msec. after the initiation of the action potential and the fibres shorten to about 57% of their rest length. Once again, the exact mechanism involved here is not entirely understood. It is this contraction of the muscle fibres which gives us the ability to move, our heart the ability to pump, and so on. And the potential which can be monitored on the skin surface during the contraction is due to the propagation of the action potential along the fibres.

2.6 THE ELECTROMYOGRAPHIC POTENTIAL. (e.m.g.)

2.6.1 HISTORY.

The relation between electricity and muscle contraction was first postulated by Luigi Galvani in 1786 with his famous frogs-legs experiment. However, initial studies of the e.m.g.s themselves were only begun about forty years ago by the neuro-physiologists, Adrian and Bronk. Their equipment was, however, very primitive and it was only after the Second World War with the improvement in electronic techniques that analysis of any real importance could be done. The first study to gain wide acceptance was that of Inman, Saunders, and Abbot in 1944, which reported their work on muscles in the shoulder region.

2.6.2 THE MOTOR UNIT.

The smooth contraction of the muscle which we observe from the outside is actually the integrated effect of the staccato-like contractions of many fibres, all firing individually, in a very random manner. Now, this random contraction occurs usually in small groups of fibres, and not each individually since each nerve axon controls a whole group of muscle fibres, not just one. The motor unit, therefore, consists essentially of the nerve cell body, the axon, its branches, and the muscle fibres that they end on. For smooth contractions, therefore, there must be complete assymetry between the firing of the different motor units. If the contractions were synchronous, as sometimes occurs, under abnormal conditions, a visible tremor results.

The number of muscle fibres per motor unit varies greatly from muscle to muscle and is dependent to a large degree on the accuracy or fineness of the activity being controlled by that muscle. Consequently, for activities such as eye movement, finger movements, and limb movements, there would be an increasing number of fibres per motor unit. Feinstein et al (1955) reported finding the number to be 108 in the first lumbrical of the hand, and 2000 in the medial head of the gastrocnemius. To get an idea of the approximate strength of a single motor unit it is interesting to note that Basmajian (1967) mentions that a slightly visible movement of a spanned joint has sometimes been noted with the stimulation of one motor unit only.

The frequency with which the motor units contract vary randomly, but there seems to be an upper limit in man of about 50 cycles/sec. This limit is obviously a physiological one and is dependent on both the time necessary for cells to recover and on the fatigue of both the muscles and the cells. Fatigue is a rather complex phenomenon and its analysis does not fall into the scope here. However, it is important to note that fatigue will decrease the frequency of firing. This frequency can to a certain degree, be consciously controlled by man, since by increasing the frequency, stronger contractions result. As the force of the contraction is increased, all the motor units increase their frequency of firing. There is apparently some inbuilt control in the order of firing of motor units to minimise fatigue.

The various fibres of each motor unit are interspaced, not lying all lumped together. Buchtal, Gould and Rosenfalck (1957) have shown that this distributed area is about 5 mm. in diameter in the human biceps.

2.6.3 ELECTRICAL CHARACTERISTICS OF THE E.M.G. SIGNAL.

The recorded e.m.g. may thus be regarded as an electrical picture of the state of contraction of the muscle or, similarly, as the electrical input to a tension transducer, the muscle. The actual relationship between the electrical signal and the output tension level is quite complex. However, Bigland and Lippold (1954) have shown that the integrated e.m.g. activity is very nearly linearly proportional to the tension exerted by the muscle. This relationship may be seen in Fig. 2.4.

140 MUSCLES ALIVE

MUSCLE MECHANICS 141

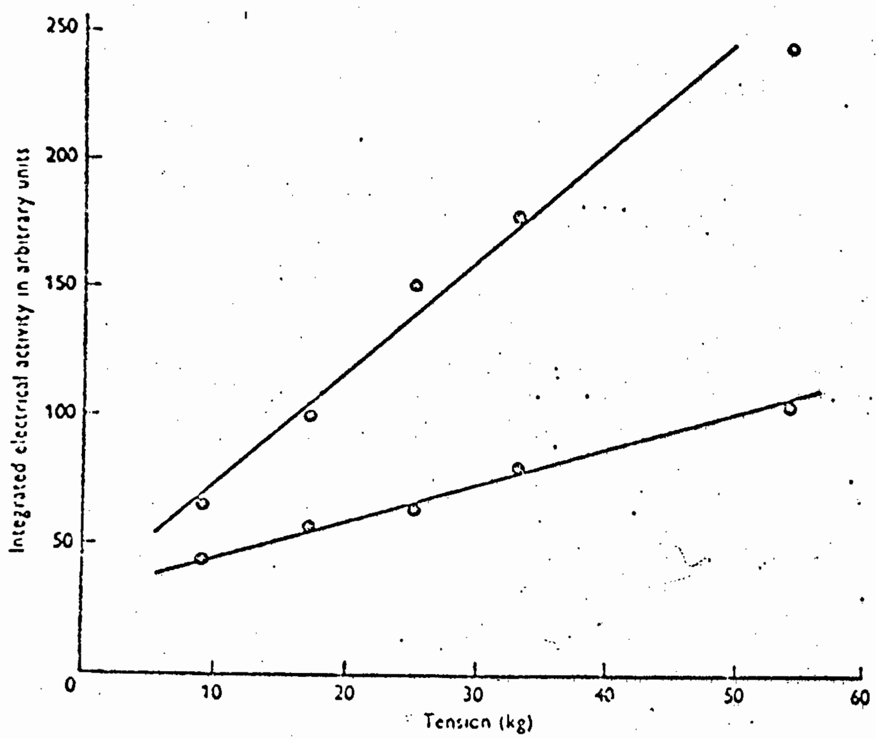


FIG. 70. The relation between integrated electrical activity (*via* surface electrodes) and tension in the human calf muscles. Shortening at constant velocity (*upper*) and lengthening at the same constant velocity. (From Bigland and Lippold, 1954a.)

Fig. 2.4

Several experimenters have tried analysing the frequency spectrum with a view to determining the number of motor units that were firing and whether they were firing synchronously or asynchronously, since this could provide an idea of the state of the muscle. A simpler method than this is the counting of the zero crossover points of the e.m.g. signal. Positive results have been obtained with this method by R.D. Fusfeld (1972). He also used this method with the first and second derivatives of the e.m.g. signal and with these results was able to determine an even better correlation between zero crossover counts and certain muscular diseases and disorders.

2.6.4 CONCLUSION.

Keeping all the above discussions in mind, it is quite obvious that the varying signal which can be monitored on the surface of the skin or (using implanted electrodes) on the surface of the muscle is just a summation of many motor unit potentials, and that through the monitoring and analysis of this signal, information can be gained both of the state of the muscle and its activity.

CHAPTER 3.

LITERATURE SURVEY OF TELEMETRY IN HUMAN GAIT STUDIES.

3.1 INTRODUCTION.

Telemetry, or the transmission of electrical signals by a radio link, is a well-established technique, finding its origins with the invention of wireless transmissions at the turn of the century. Its application to the field of Biology and Medicine was, however, very limited until after the Second World War, when, with the invention of the transistor, small battery-operated transmission systems became practically possible.

Since then, hundreds of different systems, collectively termed bio-telemetry systems, have been designed and built for transmitting various physiological parameters such as temperature, pressure, or e.c.g.s. Most of these systems, however, were designed specifically for a certain application, (1972). The publication, "Bio-medical Telemetry" by Professor R.S. Mackay, now in its second edition, has also been of tremendous assistance in the improved channelling of this knowledge.

3.2 HUMAN LOCOMOTION STUDIES.

The variables of interest in the analysis of human locomotion are many and varied. They include parameters such as the trajectories of various joints, the activity of various muscles, the relative angles between various limb portions, and so on. Of these, the most amenable to improved study with the aid of telemetry are e.m.g. potentials, foot-ground contact patterns and, perhaps in a limited way, the ground pressure profiles. Much work has been done in analysing these variables with direct wire systems. Some significant studies of this type have been: Report of Fundamental Studies of Human Locomotion, Relating to the Design of Artificial Limbs (1947); Sheffield, Gersten and Mastellone, (1956); Milner, Quanbury and Basmajian, (1970).

3.3 E.M.G. TRANSMISSION.

The first attempt, however, at the actual telemetering of e.m.g. signals, was the system of Battye and Joseph (1965). This system consisted of a V.H.F. transmitter with a single F.M. channel. The frequency bandwidth of the single channel was

55 Hz. to 4.5 KHz., the authors having felt this to be sufficient for e.m.g. analysis. The system had a range of 50 metres in air. It was used by Battye and Joseph for monitoring the following muscles:- tibialis anterior, different portions of the quadriceps muscle group, the hip flexors and the gluteus medius and maximus. The great draw-back of their system was the inability for simultaneous monitoring of several muscles. Another system designed for monitoring e.m.g. potentials is that of Hirsch, Kaiser and Petersen, (1966). The monitored e.m.g.s here were used for controlling a prosthesis and not for analysis, and hence an implant was used. This system was a passive system having the necessary power transmitted to an implanted receiver and the e.m.g. transmitted out as a modulation on a different frequency.

Other systems have been designed since the one described above by Battye and Joseph, in which several channels were multiplexed. These systems were, however, not specifically designed for e.m.g. transmission and the channels had only sufficient bandwidth to accommodate part of the wide band of e.m.g. signals. Notable among these are that of H. Fischler, N. Peled and S. Yerushalmi, (1967), who constructed a six-channel F.M./F.M. system with signal channel bandwidths of 500 Hz.; and R.D. Robrock 11, W.H. Ko, (1967), whose F.M./F.M.

system also had signal channels each with bandwidths of approximately 500 Hz.

A multi-channel system specifically designed with large channel bandwidths required for e.m.g. analysis has, however, not been designed, and would hence be a useful next step.

3.4 FOOT-GROUND CONTACT TRANSMISSION.

Foot-ground contact is an extremely important parameter to monitor in order that the e.m.g.s of the various muscles may be referenced to the swing and stance phases of the walking pattern. In direct line systems, metal walkways with conductive strips attached to the shoes have been used by many groups, viz., Schwartz et al, (1934), Smith et al, (1960), and Milner (1970). Using a similar method, variable capacitive walkways have been designed by Jones et al, (1966). These, however, do not lend themselves to the ease of portability offered by telemetry. Battye and Joseph in their system made use of photographic techniques to determine foot-ground contact, but this method has a rather limited use and does not yield the information in a readily usable form.

A compact transmitter oscillator was designed by Herron et al, (1967). This could be placed in the heel of a shoe and ground contact was indicated by the presence or absence of the transmitted signal.

An improvement on this system has just been perfected by Winter et al, (1972) in which a whole array of micro-switches were embedded in the sole of a shoe. These were arranged together with a resistor network to give varying output voltage levels and the information was transmitted by means of an F.M./F.M. telemetering system. This system not only permitted referencing foot-ground contact, but also the transfer of weight over the sole surface.

3.5 PRESSURE TRANSMISSION.

Schwartz and Heath (1947) developed a system using pressure sensitive transducers to monitor the pressure exerted on certain parts of the foot. Its application is not so significant due to the complexity of forces acting on the sole face and the difficulty of measuring this with isolated transducers in the sole surface. Many researchers have used force plates to monitor this complex pressure pfofile. There is, however, a commercially available system, the Stanmore Sandal, which utilises sandals fitted with load cells, which enable the total foot force (vertical) to be telemetered.

3.6 CONCLUSION.

Several systems have been mentioned in which telemetry was used to aid in locomotion studies. The scope of bio-telemetry in human locomotion is not so limited though, and as the importance of obtaining unrestrained information increases, so will the use of bio-telemetry.

CHAPTER 4.

WHY TELEMETER E.M.G.S.?

4.1 INTRODUCTION.

When telemetry is used to replace a set of wire leads, a very simple system is immediately transformed into an extremely complex one. Consequently, it is important to evaluate the expense involved and compare it with the advantages gained before embarking on the change-over. This summing-up is especially important in the cases where systems have already been designed without telemetry and useful results obtained. The monitoring of e.m.g.s is just such a case.

4.2 DISADVANTAGES OF TELEMETRY.

The disadvanteges of telemetry are of two main types, expense and inaccuracy.

It is obvious that the replacement of a straight set of wires by a multiplexed transmission system will require far greater capital outlay. Firstly in the cost of the circuit components, and then in the time necessary to design, build and develop the telemetry system.

Due to this increased complexity and increased number of components, greater possibilities exist for error and malfunction, since more electronic components can fail and errors can accumulate from section to section. The performance of the system is thus only as good as the worst segment.

A further difficulty exists in the fact that transmission fields are not evenly distributed in the area around the transmitter. The field strength drops off rapidly as the distance from the transmitter increases, and steps have to be taken that this variation is not misinterpreted as a variation in the information signal. The presence of some shielding material, e.g. a brick wall, will also affect the strength and could lead to a similar misinterpretation.

Finally, the far field strength may vary with the orientation of the receiver. This is due to the fact that the transmission field is not circularly uniform and the field strength may tend to zero at certain angular portions of the field. Such problems do not exist with direct wire systems.

It is obvious therefore that one is bound to encounter far greater technical difficulties and these have to be weighed against the improved information content.

4.3 ADVANTAGES OF TELEMETRY.

4.3.1 GENERAL ADVANTAGES OF BIO-TELEMETRY.

It is of interest to note how telemetry has been used to advantage in other spheres of medicine and biology. Perhaps the most important advantage offered by telemetry is the increased mobility offered to the patient or animal being monitored. If, for example, the purpose of the monitoring is the collection of data for an experimental study, the results have greater validity if the subject is unrestrained. In the case of animals, this is even more of a reason, since they could very easily get tangled up in the long leads.

In the case of hospital monitoring of the e.c.g., temperature, etc., of patients in intensive care wards, telemetry not only increases the patient's mobility, but improves his comfort and general frame of mind, since the absence of long leads makes him less aware of the equipment connected to him.

Small pill-size radio transmitters or endo-radiosondes, which can be swallowed without difficulty, provide easy access to measurement in regions of the body otherwise inaccessible without the use of surgery. However, when surgery is necessary to be able to monitor certain variables, a transmission system would obviate the necessity of wires through the skin which would undoubtedly increase the danger of infection. Such a case is the transmission of an external stimulus pulse to the bladder muscle, to assist incontinent people to pass water. Surgery is also of advantage sometimes for implantation of electrodes in animals to stop them scratching or pulling them off. Here again, telemetry is invaluable.

4.3.2 ADVANTAGES OF TELEMETERING E.M.G.S.

A large number of systems have been designed and used extensively for monitoring e.m.g.s with long monitoring leads and significant information has been obtained from these systems. However, the presence of long wire leads inevitably inhibits the walking pattern of the subject, thereby creating artefacts in the collected information. Since these results will be used both for clinical diagnosis and in the improved design of artificial limbs and muscle stimulus systems, the information must be extremely accurate, and the removal of the artefact might be of significant importance. Telemetry would noticeably reduce this artefact.

Increased mobility is also of importance in e.m.g. measurements, especially in more general e.m.g. studies such as during sports or any other activities where freedom of movement beyond a restricted walkway and without encumbering leads is necessary.

Another important reason for telemetering e,m.g.s is the increased mobility offered to the experimenter, since telemetry systems are far more portable. Measurements could be made in crowded department stores or out in the street with the minimum of difficulties in transporting the equipment.

4.4 CONCLUSION.

When considering the above facts, it is obvious that telemetering e.m.g.s offers a definite advantage over direct wire systems and that the use of telemetry will provide for the collection of data essential to a complete understanding of the locomotion process.

CHAPTER 5.

SYSTEM DESIGN.

5.1 BASIC SYSTEM REQUIREMENTS.

As an initial step, it was essential to lay down the basic specifications around which the system was to be designed.

5.1.1 CHANNEL SPECIFICATIONS.

It is quite obvious that a multi-channel e.m.g. system would be convenient for any human locomotion studies, since the phasic activity of the muscles could be determined far easier than by monitoring muscles one at a time. Four e.m.g. channels were felt to be sufficient, as thereby the four main muscle groupings (tibialis anterior, gastrocnemius, hamstrings and quadriceps) could be studied simultaneously on one leg. Although increasing the number of e.m.g. channels much above four would decrease the advantage of free movement gained by using a telemetry system, due to the increased bulk to be carried by the subject, a decrease in the size of the differential amplifiers and sub-modulators would permit increasing this number.

In addition to the e.m.g. channels, one or two footswitch channels were needed to indicate toe-ground, heel-ground, and flat foot contact. This would facilitate a means of referencing the phasic activity of the muscles.

The bandwidth of an e.m.g. signal, as mentioned in Chapter 2, is extremely broad, having frequency components up to and beyond 10 KHz. The higher energy components, however, extend to just over 500 Hz. with the energy content at higher frequencies being small. Since the total bandwidth of a telemetry system is necessarily limited, and increasing the channel bandwidth would obviously limit the number of channels, a channel bandwidth of 2 KHz was chosen as being an optimum value. This permitted a sufficient amount of e.m.g. information to be transmitted and, as will be seen later, permitted the required number of channels to be accommodated.

5.1.2 SIZE AND WEIGHT SPECIFICATIONS.

Since one of the primary advantages of this telemetry system is the freedom of movement that it will give to the patient or the subject being monitored, it was essential that the size and weight of that part of the system to be carried by the individual (viz. the transmission system) be made as small as possible. Consequently, any complexity was to be, where possible, incorporated into the receiving end of the system, which is remote from the subject. A reasonable limit to the weight of the entire apparatus carried by the patient would be about 1 kg. for both adults and children, and the size of any individual section should not be greater than 5 cm. x 5 cm. x 8 cm.

Each e.m.g. amplifier and sub-modulator was to be self-contained, with the exception of the central power supply. This was important since the actual electrode leads need to be as short as possible in order to minimise both movement arte-facts and noise that enter the system. This required then that the e.m.g. amplifier be located near the recording site. A sturdy casing and once again minimum size were thus important for these amplifiers.

The power supply, the transmitter itself and its aerial were to be mounted on a belt around the person's waist, as this was the position where it was least likely to cause any inconvenience. The footswitch oscillators could be mounted on the waist as well, though if sufficient miniaturisation was possible, they could have been embedded in the shoe sole. This, however, would not have permitted simple interchangeability of footwear and so was not considered. The switches to indicate toe and heel contact were adhered to the sole surface. This required them to be flat and not much greater than 3 cm^2. This was once again preferable to having the switches embedded in the sole because of the ease of changing shoes from subject to subject.

Further miniaturisation of the amplifiers and sub-modulators is possible with thick film technique.

5.1.3 RANGE SPECIFICATION.

The range of the transmission system should be as large as possible without creating too great a power drain on the batteries. Since many measurements will be made in an external environment, such as on a sports field or on the street, the minimum range of the system should be approximately 50 metres in the open air.

5.2 SYSTEM LAYOUT.

Keeping the above criteria in mind the system layout of Fig. 5.1 was drawn up.

Heel and toe contact were indicated by closing the relevant switches, thereby turning on the oscillators. These oscillators, operating at different frequencies, then frequency modulated (F.M.) the transmitter.

The e.m.g. signals monitored with surface electrodes are of the order of 1 mV. and were consequently amplified 100 to 200 times so that they could modulate the sub-carrier oscillators. The use of differential amplifiers here was essential due to the presence of 50 Hz. mains interference. To satisfactorily reduce this noise, a common mode rejection ratio in excess of 90 db. was needed.

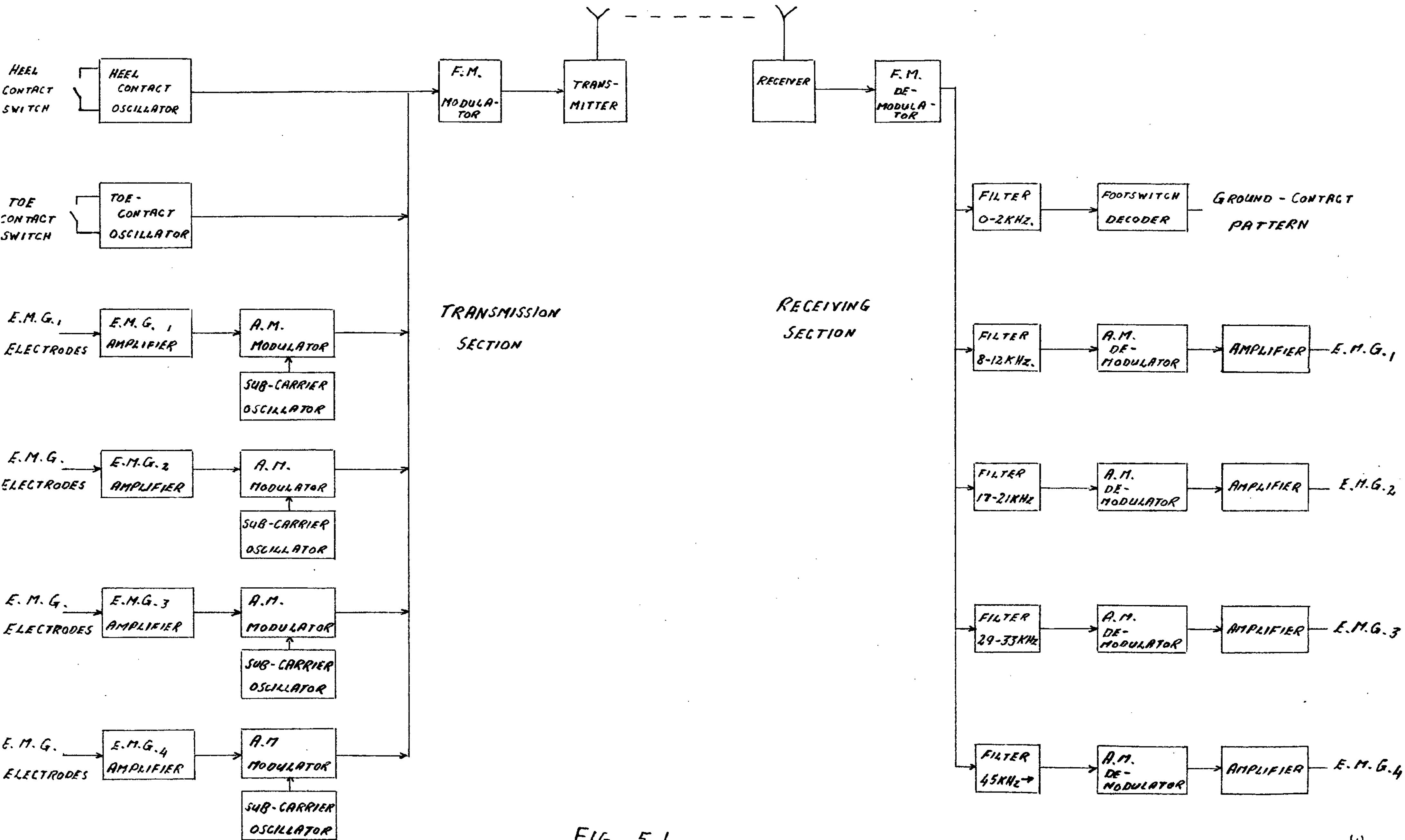

HEEL CONTACT SWITCH
HEEL CONTACT OSCILLATOR
TOE CONTACT SWITCH
TOE-CONTACT OSCILLATOR
E.M.G.1 ELECTRODES
E.M.G.1 AMPLIFIER
A.M. MODULATOR
SUB-CARRIER OSCILLATOR
E.M.G. ELECTRODES
E.M.G.2 AMPLIFIER
A.M. MODULATOR
SUB-CARRIER OSCILLATOR
E.M.G. ELECTRODES
E.M.G.3 AMPLIFIER
A.M. MODULATOR
SUB-CARRIER OSCILLATOR
E.M.G. ELECTRODES
E.M.G.4 AMPLIFIER
A.M MODULATOR
SUB-CARRIER OSCILLATOR
F.M. MODULATOR
TRANSMITTER
TRANSMISSION SECTION
RECEIVER
F.M. DE-MODULATOR
RECEIVING SECTION
FILTER 0-2KHz.
FOOTSWITCH DECODER
GROUND-CONTACT PATTERN
FILTER 8-12KHz.
A.M. DE-MODULATOR
AMPLIFIER
E.M.G.1
FILTER 17-21KHz
A.M. DE-MODULATOR
AMPLIFIER
E.M.G.2
FILTER 29-33KHz
A.M. DE-MODULATOR
AMPLIFIER
E.M.G.3
FILTER 45KHz→
A.M. DE-MODULATOR
AMPLIFIER
E.M.G.4

FIG. 5.1

The amplified e.m.g. signal was then used to amplitude modulate (A.M.) a sub-carrier oscillator. This creates a frequency shift of the e.m.g. signals allowing them all to frequency modulate the transmitter in a different frequency band. An important reason for choosing A.M. sub-carrier modulation was that the circuitry is far simpler than that of F.M. The technical reasons will be discussed in paragraph 5.3.

The F.M. transmitter with carrier frequency of 108 MHz. then transmitted its signal which was received by an F.M. receiver. A ratio detector was used to separate the sub-carrier modulated frequencies from the transmission frequency. Each channel was then separated with the aid of filters, A.M. demodulated and amplified before being fed into the recording apparatus.

The de-coding of the footswitch signals was achieved as follows: the two oscillator frequencies were separated from each other by additional filters. They were then integrated and converted to a set voltage to indicate the presence or absence of the oscillating frequency and hence the presence or absence of toe and heel ground contact.

5.3 MULTIPLEXING.

If each individual e.m.g. was given its own transmitter with a specific frequency band, this would have required a tremendous amount of duplication and bulk, both in the equipment carried by the monitored individual and in that at the receiving end. Consequently, multiplexing the channels was essential to the viability of the system. Two types of multiplexing are available: frequency division multiplexing and time division multiplexing.

5.3.1 FREQUENCY DIVISION MULTIPLEXING.

5.3.1.1 FREQUENCY MODULATION.

Considering initially frequency division multiplexing: This requires either amplitude modulating or frequency modulating a sub-carrier oscillator and using these in turn to modulate the transmitter oscillator. Using F.M. sub-modulation, the newly created frequency components are:

$$\phi_{FM}(t) = A \sum_{n=-\infty}^{\infty} J_n(m_f) \cos(\omega_c + n\omega_m)t$$

where A is a constant multiplying factor,

$J_n(m_f)$ is a complex integral,

ω_c is the carrier frequency,

and ω_m is the modulating frequency.

Without going into any great detail with the above equation, it can be seen that a whole string of frequency components has been created, viz., $(\omega_c \pm \omega_m)$, $(\omega_c \pm 2\omega_m)$, $(\omega_c \pm 3\omega_m)$ etc. Where ω_c is the sub-carrier frequency and ω_m is the e.m.g. frequency.

Thus in addition to the desired side-bands $(\omega_c \pm \omega_m)$ a whole string of secondary side-bands of decreasing amplitude are created as well. If these were not filtered out, before mixing all the frequencies at the transmission end, they would interfere with the other channels. Thus separation of the channels at the receiving end would be extremely difficult and crosstalk would result. Consequently, filters would be required on the transmission side, increasing the bulk carried by the subject.

5.3.1.2 AMPLITUDE MODULATION.

Using A.M. the following results:
If the Fourier transform of $f(t)$ is $F(\omega)$
then the Fourier transform of
$f(t)\cos\omega_c t$ is $\frac{1}{2}(F(\omega+\omega_c)+F(\omega-\omega_c))$
where $f(t)$ is the e.m.g. signal
and $\cos\omega_c t$ the sub-carrier oscillator signal.

Thus only two side-band frequencies are created. Also, since the transmission system istself was frequency modulated no extra noise rejection could be gained by using F.M. in the sub-carrier and so A.M. was chosen.

The real problem however, was the fitting of the desired number of channels into the system bandwidth of 50 KHz. , which was dictated by the bandwidth limitation of the receiver. This will be discussed further in Chapter 6. Fig. 5.2 shows how the channel frequencies could be spaced out.

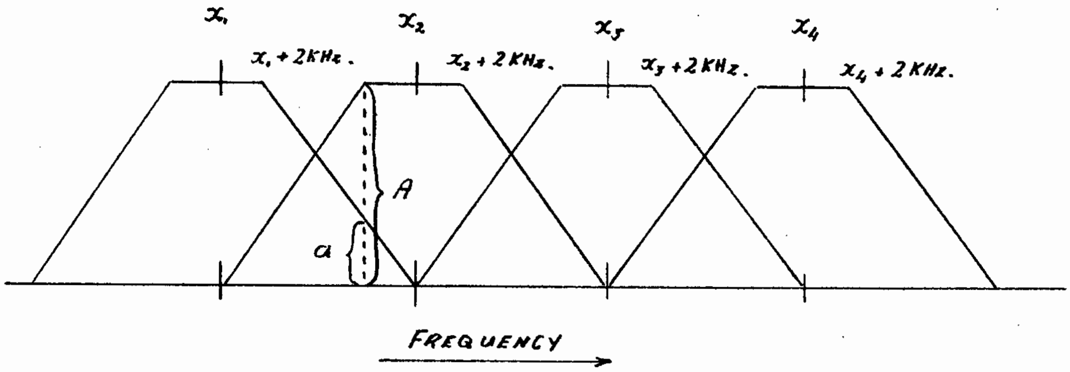


Fig. 5.2

"a" is a measure of the interference between two channels. 5% was chosen as the maximum acceptable ratio of a/A. Thus in order to be able to separate out the channels at the receiving end, filters were required to attenuate by z db. within a set frequency.

where $z = 20 \log V_2/V_1 = 20 \log .05$

$= 25.82$ db.

The minimum frequency for x_1 must be much greater than the 2 KHz. modulating signal for efficient de-modulating, and so x_1 was chosen equal to 10 KHz. Keeping all the channels within the 50 KHz. band-width, the other carrier frequencies were chosen at:

$x_2 = 18$ KHz.

$x_3 = 30$ KHz.

$x_4 = 47$ KHz.

Fitting these values to Fig. 5.2 it can be seen that filters capable of 25 db. attenuation within a third of an octave were required.

Single side-band modulation would have

permitted less stringent filter requirements but would increase the complexity and size of the transmission side, since filtering of the amplitude modulated signal would be required before frequency modulating the transmitted signal. And so, once again, this was not considered. Frequency division multiplexing was chosen over time division multiplexing; the difficulties are mentioned below.

5.3.2 TIME DIVISION MULTIPLEXING.

Time division multiplexing is accomplished by discreet sampling of the information in the individual channels and the staggering of these samples in time before using them to modulate the transmitter. Any deviation from ideal sampling negatively influences the quality of the reproduced signal.

From the sampling theorem,the sampling rate must be twice the size of the sampled frequency, i.e. 4 KHz. Now, since four channels were being interwoven, the real pulse frequency required was: 16 KHz. However, the real problem is that to satisfy the ideal sampling criteria, the samples must be as near square pulses as possible, and this means, practically speaking, the inclusion of up to the tenth harmonic of the pulse. This would necessitate a bandwidth of approximately 200 KHz. for the transmission of 4 channesl. The bandwidth of 50 KHz. would thus have been prohibitive and therefore time division multiplexing was not used.

Time division multiplexing also requires bulky encoding circuitry which might increase the size of the system carried by the patient by about 50%. This was a further point against its use.

5.4 TRANSMISSION SYSTEM.

Frequency modulation for the transmission system has several advantages over that of amplitude modulation. The most significant being that in the case of A.M., should a subject, in walking, pass behind some obstructing medium like a pillar, etc., this would decrease the power picked up by the receiver. This would be interpreted as a decrease in the amplitude of the transmitted signal and hence, an inaccuracy would result in the information received. Whereas in the case of frequency modulation it is the variation of frequency which contains the information and this is not affected by any obstructing media. In addition, the inherently better signal to noise ratio of F.M. compared to A.M. was taken into account in this choice, especially due to the very low level of transmitted signal power.

The use of very high frequencies permitted the use of small aerials, whereas if lower frequencies were used, a complicated aerial system hung in a loop around the recording room would have been necessary in order to monitor the very small signal power of the receiver. This would subtract one of the main advantages of the system, which is portability. Another reason for using very high frequencies is that commercial F.M. receivers use the frequency band 88-108 MHz., and so a cheap F.M. receiver could be used for reception. The actual frequency chosen was 108 MHz., since this is right on one end of the band and as such would not be blacked out by regular broadcasting stations.

CHAPTER 6.

ELECTRONIC CIRCUIT DESIGN.

6.1 THE TRANSMITTER.

6.1.1 THE FOOTSWITCH SECTION.

6.1.1.1 THE OSCILLATORS.

Two signals were required: one to indicate toe-ground contact and the other heel-ground contact. For this purpose, two low frequency, parallel T, phase-shift oscillators were used, diagrams of which may be seen in Fig. 6.1.

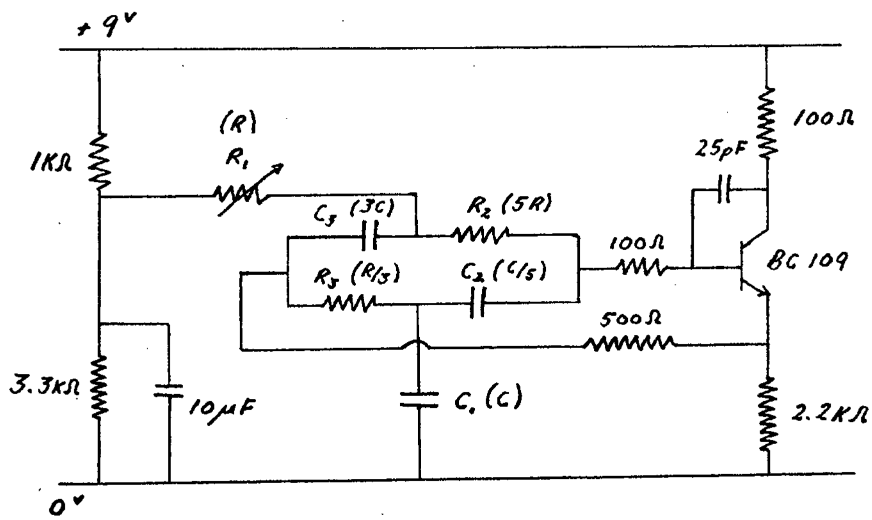


Fig. 6.1

The frequency of oscillation:

$$f = \frac{1}{2\Pi\ RC}$$

Having decided values for R and C and consequently for R_1 R_2 R_3 C_1 C_2 C_3 it was found advisable to make R_1 variable in order to set the frequency accurately and to keep distortion of the signal output down to a minimum. The 500Ω feedback resistor could be replaced by a miniature bulb of 24v., 40mA. to provide for temperature compensation, lack of which causes frequency drift and signal distortion of the generated sine wave. This was, however, not found to be a problem here, although bulbs were used in further oscillators where activity was critical.

The chosen frequencies for the two oscillators were 400 Hz. and 2 KHz. The large separation between the two frequencies allowed the use of simple filtering at the receiving end without any crosstalk between the toe and heel signals. Sine wave oscillators were necessary here since any other type of signal would contain harmonics of such frequencies that avoidance of crosstalk would have been very difficult.

For the 400 Hz. oscillator:-

$$f = 400 = \frac{1}{2\pi RC}$$

Choose $R = 3K\Omega$ $\quad\therefore\quad C = \dfrac{1}{2\pi x 4 x 3 x 10^{5}}$

$= .133\mu F$

$\therefore R_1 = R = 3K\Omega$ $\qquad C_1 = C = .133\mu F$

$R_2 = 5R = 15K\Omega$ $\qquad C_2 = C/5 = .027\mu F$

$R_3 = R/3 = 1K\Omega$ $\qquad C_3 = 3C = .39\mu F$

For the 2 KHz. oscillator:-

$R_1 = R = 3K\Omega$ $\qquad C_1 = C = .0266\mu F$

$R_2 = 5R = 15K\Omega$ $\qquad C_2 = C/5 = .0053\mu F$

$R_3 = R/3 = 1K\Omega$ $\qquad C_3 = 3C = .0798\mu F$

6.1.1.2 THE FOOTSWITCHES.

The footswitch is required to connect the oscillator to the F.M. transmitter. The ideal place to put it then, would have been between the output of the oscillator and the input to the transmitter, as this would eliminate any transients in the switching ON and OFF of the oscillator circuit. However, this would then require the oscillators to draw current continuously - a rather undesirable situation. It was also found that the oscillations filtered through the transmitter along the power rails, bypassing the output switch. Consequently, the switches were used to switch the positive supply to the circuit. The transients were found not to be prohibitive.

Two designs of switches were made and tested.

a) Brass shim switch.

This switch consisted of a piece of thin cardboard sandwiched between two pieces of brass shim. A hole was made in the centre of the cardboard through which the two pieces of shim could touch. A small coin of approximately 1.5 cm. was stuck to one of the brass shims to decrease the pressure required to close the switch. (The hole in the cardboard was just a little bigger than the small coin). The overall size of the switch was approximately 4-5 cm. square and the thickness was about .2 cm.

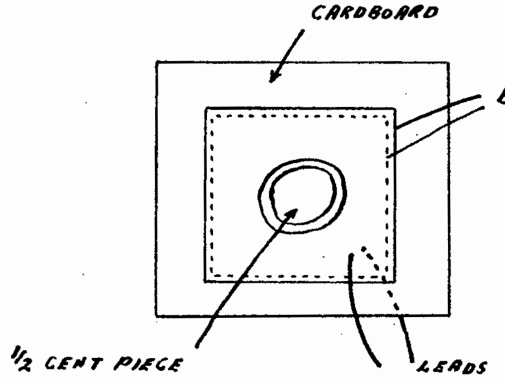


Fig. 6.2

The switch performed well under quite tough conditions. Although after a while, the shim began to bend, lose its springiness and the performance of the switch deteriorated. This was due to the shim beginning to take the shape of the small coin. Thus a second switch was tried.

b) <u>Rubber tubing switch</u>.

This switch consisted of a short piece of .5 cm. rubber tubing about 3 cm. long, with two pieces of steel wire piercing the tube so that they could make contact at one point inside the tube when the tube was compressed. Talcum powder was sprinkled inside the tube to prevent it sticking closed after the pressure had been released.

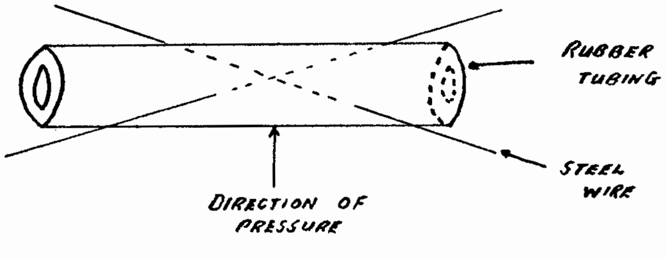


Fig. 6.3

This switch had a better long-term wear than the previous one. It was less erratic in its behaviour and could be made noticeably smaller. Consequently, this was the switch finally used.

Both these switches satisfied the initial criteria laid down in the previous chapter, viz., they may be applied to any shoe without any major alterations to the sole. An excellent switch which, however, requires a special hole in the sole is that described by Winter et al (1972) of the University of Manitoba, Canada. In this case, five of these switches were embedded in the sole of one shoe.

6.1.2 THE E.M.G. SECTION.

6.1.2.1 ELECTRODES.

The electrodes used for monitoring the e.m.g. were the commercially available "Disposel"[1] electrode kits. Before applying the electrodes, the skin was first rubbed with an abrasive electrode paste, "Redux", to remove the top dead skin layer, washed, and then rubbed with alcohol to dissolve any oils and fats which might short circuit part of the signal.

[1] These electrodes are made from silver/silver chloride with adhesive already attached, are supplied in disposable kits together with Trucon electrode paste. These kits are available from Electrodyne.

The electrodes were then filled with the electrode paste supplied in the kit and applied when the alcohol had dried.

6.1.2.2 AMPLIFIERS.

The amplification of e.m.g. signals requires a high gain differential amplifier. The common mode rejection ratio should be of the order of 90-100 db. to suppress 50 Hz., and other noise. Three alternative differential amplifier designs were investigated.

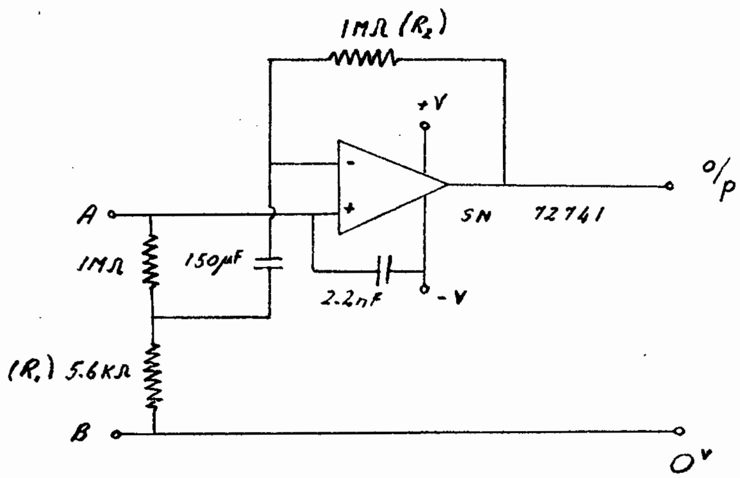


Fig. 6.4

a) The above differential amplifier has the enormous advantage of small size, since only one operational amplifier, the commercially available SN 72741, is required. This circuit has a C.M.R.R. well in excess of 90 db.

However, since the two inputs are not completely isolated from earth, if more e.m.g.

amplifiers are coupled to the same earth, a certain amount of crosstalk results. This is due to the fact that the amplifier will register the signal level of the A electrode, not only with respect to its B electrode, but with respect to all the other B electrodes as well. An experiment was performed to determine this crosstalk (this experiment appears as Appendix A) and it was found to be about 2% per channel. The amount of crosstalk increased as the linear function of the number of channels with the same earth. The circuit was consequently felt to be unreliable, since the admittance of an inevitable 6% crosstalk initially would completely upset the system. Its use, however, is suggested if only one or two variables are being monitored.

Amplification is determined by the ratio

$$R_2/R_1 = 1\text{M}\Omega/5.6\text{K}\Omega$$
$$= 178.5$$

b) The answer to the problem of the above differential amplifier was to have floating inputs. This may be seen in Fig. 6.5.

The amplifier consists of two high input impedance buffers feeding into an SN 72741 connected as a differential amplifier. In this circuit SN 72741 were used for the buffer stages, but improved performances can be acquired using LM 310 instead. The C.M.R.R. here was also in excess of 90 db. and crosstalk less than .1%.

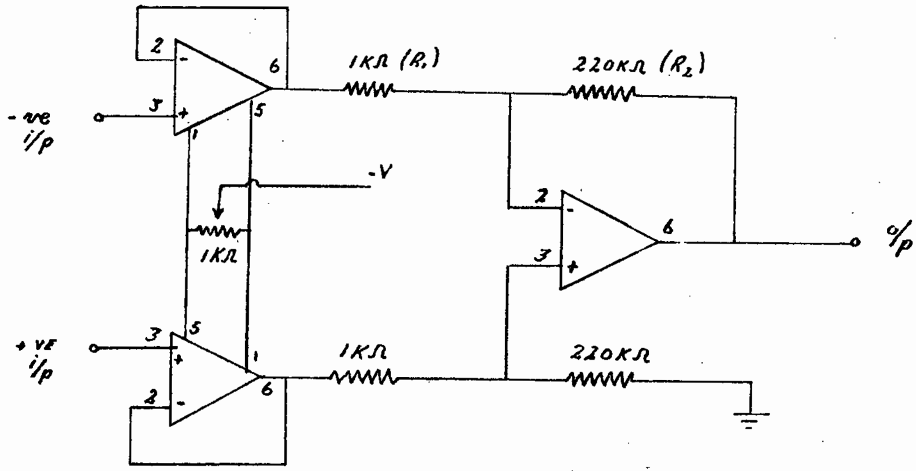


Fig. 6.5

Amplification is determined by the ratio

$$R_2/R_1 = 220K\Omega/1K\Omega = 220$$

c) When miniaturising the circuitry was considered, the following circuit was found. With the use of an SN 72558, which is a dual op-amp in an 8-pin package, the size of this differential amplifier is one third that of the one in case b.

This circuit has two main disadvantages which may, however, be overcome in this application. Firstly, the output has a d.c. offset, but since the coupling used was a.c., this was not a problem. Secondly, the resistors must be within 1% of the specified values, otherwise the C.M.R.R. is greatly reduced.

This circuit is shown in Fig. 6.6.

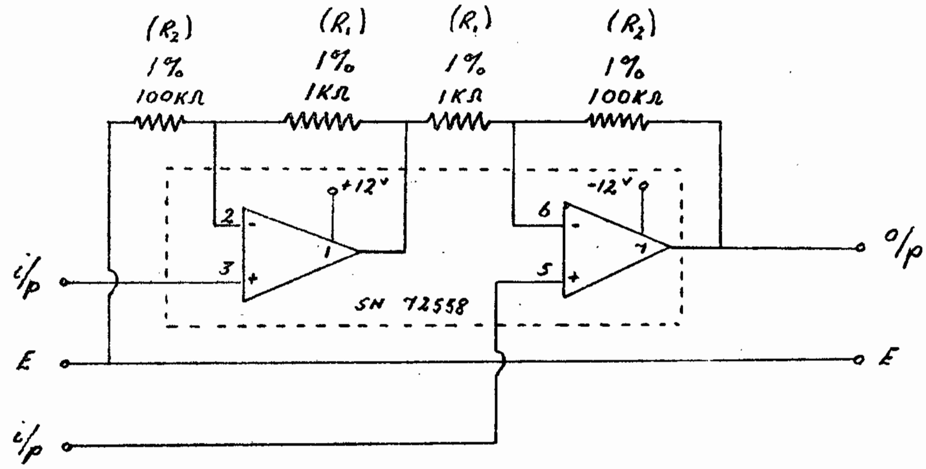


Fig. 6.6

The amplification in this circuit is given by

$$R_2/R_1 = 100K\Omega/1K\Omega$$
$$= 100$$

6.1.2.3 SUB-CARRIER OSCILLATORS.

The oscillators used here were once again phase shift, sine wave oscillators, with the same electronic design as those used for the footswitches. Here, however, 24v, 40mA bulbs were used as the feedback element to give temperature compensation since the frequencies used here have critical values.

For the reasons illustrated in Chapter 5, these sub-carrier oscillators were amplitude modulated by the e.m.g. signals. This was accomplished by replacing the emmitter resistor in the oscillator circuit with a transistor. The transistor was biased by means of resistors to the centre of its active region and the output of the e.m.g. amplifier was a.c. coupled to its base. The variation of the e.m.g. amplitude thus causes a variation in the collector-emitter resistance, which is a variation of the emitter resistor of the oscillator. This in turn varies the feedback current, which changes the output voltage accordingly. The transistor should be biased in the centre of its active region to ensure a linear variation of resistance with e.m.g. voltage.

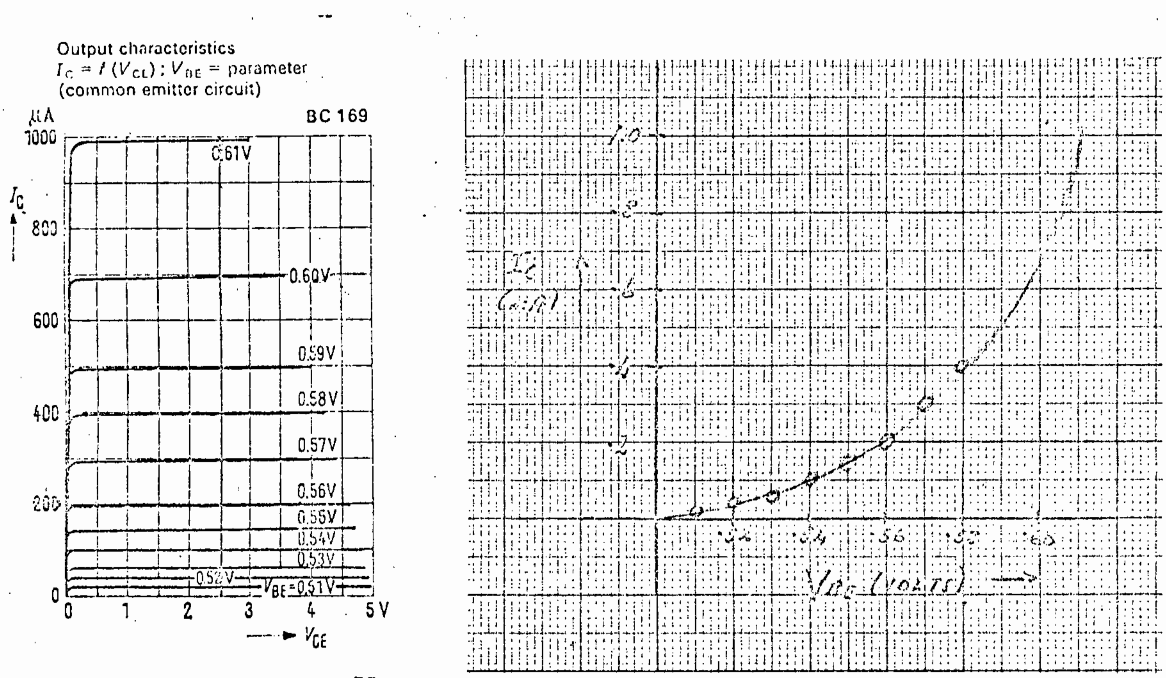


Fig. 6.7

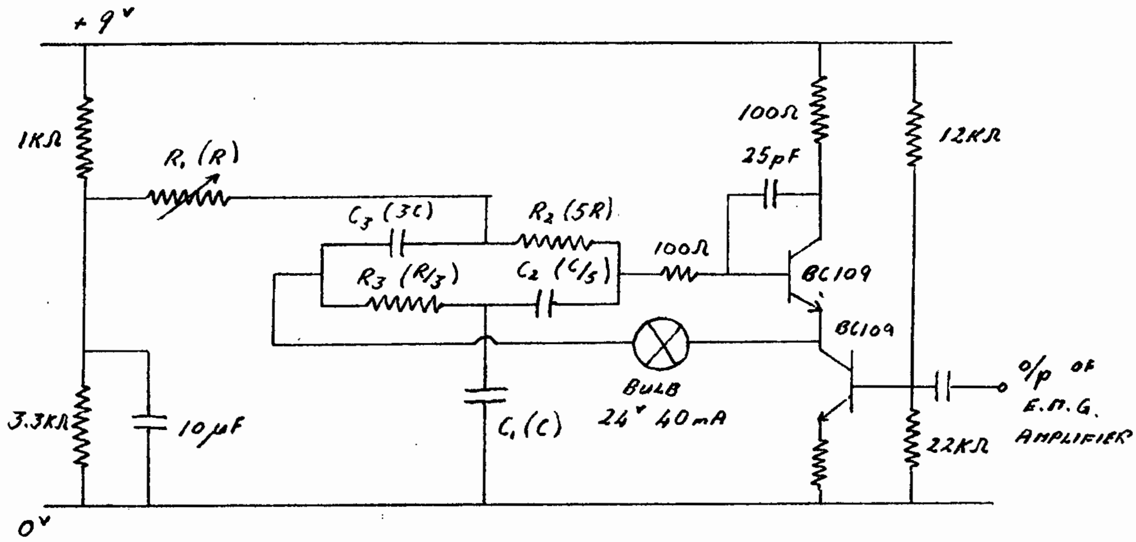


Fig. 6.8

The following are the calculated component values for the four sub-carrier oscillators at the frequencies decided upon in Chapter 5. The method of calculation is the same as that used in 6.1.1.1.

For 10 KHz:-

R_1 = 3KΩ	C_1 = .0053 μF
R_2 = 15KΩ	C_2 = .0011 μF
R_3 = 1KΩ	C_3 = .0159 μF

For 19 KHz:-

R_1 = 3KΩ	C_1 = .0028 μF
R_2 = 15KΩ	C_2 = .00056 μF
R_3 = 1KΩ	C_3 = .0084 μF

For 31 KHz:-

R_1 = 3KΩ	C_1 = .00171 μF
R_2 = 15KΩ	C_2 = 342 pF
R_3 = 1KΩ	C_3 = .00513 μF

For 47 KHz:-

R_1 =	3KΩ	C_1 =	.00113 μF
R_2 =	15KΩ	C_2 =	226 pF
R_3 =	1KΩ	C_3 =	.00339 μF

6.1.3 THE TRANSMITTER SECTION.

The transmitter section consists of a frequency modulated 108 MHz. transmitter. An output stage was added to increase stability.

6.1.3.1 THE TRANSMITTER OSCILLATOR.

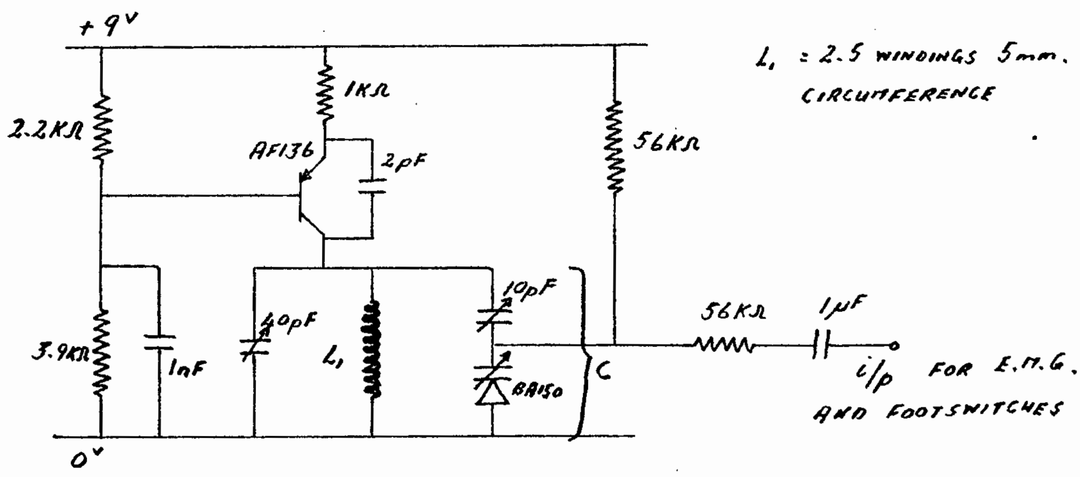


Fig. 6.9

The transmitter consisted simply of a slightly modified Colpitts oscillator. The reason for the modification is that in the present configuration, the capacitance of the tank circuit can be easily altered, thus changing the frequency of oscillation of the circuit. This permitted frequency modulation of the circuit.

The frequency of oscillation is given by:

$$f = \frac{1}{2\Pi\sqrt{L_1 C}}$$

For f = 108 MHz C = 40pF

$$L_1 = .378\mu H$$

This extremely low value of inductance is rather difficult to construct with any accuracy. By winding 2.5 turns of gauge 18 wire on a 5 mm. circumference former this value is achieved as closely as possible and further tuning is effected by varying the capacitance.

6.1.3.2 THE MODULATOR.

By using the variable capacitance of a reversed biased diode BA 150 in series with a 10 pF capacitor and this in parallel with the tank circuit of the oscillator, the frequency of oscillation was made dependent on the voltage applied to the varicap. The a.m. e.m.g. signals and the footswitch signals were then a.c. coupled to this point as shown in Fig. 6.9, and they consequently frequency modulated the transmitter oscillator. This modulator has a bandwidth of 50 KHz. as shown on the graph 6.1. It is this modulator that has the smallest bandwidth in the whole system, but it fits into the specifications laid down in the previous chapter.

6.1.3.3 THE OUTPUT SECTION.

The frequency of oscillation of the oscillator is not very stable and may vary quite dramatically as its capacitive coupling with earth is altered. For this reason, a buffer output section was added. With the buffer added, stability increases and the oscillating frequency is unaffected by variations in capacitive coupling.

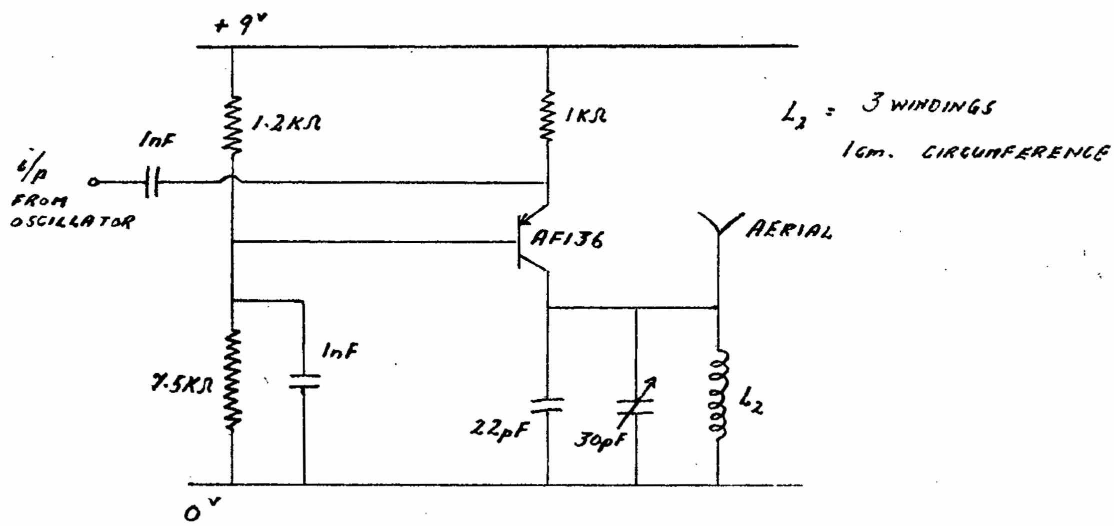


Fig. 6.10

6.1.4 LOADING OF OSCILLATORS.

The outputs of all the five channels are connected directly to the one input of the F.M. transmitter. To prevent any substantial e.m.g. information being fed onto one of the other channels, modulating it and thereby creating crosstalk, the oscillators were loaded as in Fig. 6.11. Unfortunately, a certain amount of output voltage is lost across the 9KΩ resistors, but this is not prohibitive.

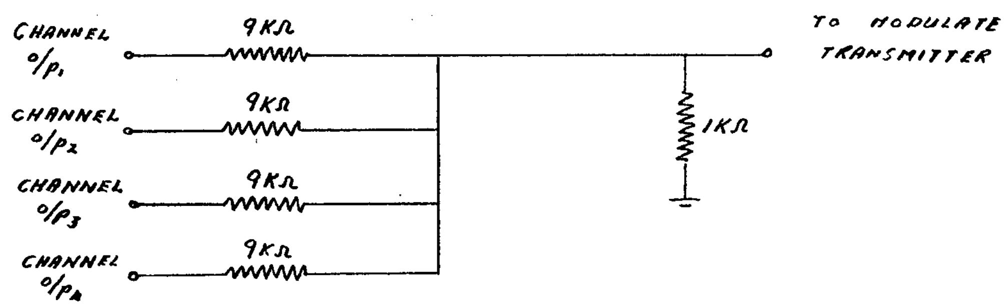


Fig. 6.11

6.1.5 POWER SUPPLY.

Miniature 9v. Eveready batteries[1] were used. Two batteries were used for the positive and negative supplies of the op-amps in the differential amplifiers. A separate battery was used for the transmitter itself so as to further stabilise its oscillating frequency. The whole circuit has a low current drain and one set of batteries lasts two weeks with about five hours' use every day.

6.1.6 AERIAL.

A short copper rod was attached to the output of the transmitter buffer to act as an aerial. Using a commercial receiver, the signal could be picked up to a range of 30 metres.

Fig. 6.12 is a photograph of the transmitting apparatus.

[1] These batteries were PM3 Eveready with size equal to 4.5 x 2 x 1.5 cm.

Fig. 6.12

6.2 THE RECEIVING SECTION.

6.2.1 THE RECEIVER.

Since the frequency of transmission was chosen as 108 MHz., i.e., within the range of the commercial F.M. receiver, an ordinary F.M. receiver could be used. The actual receiver used was an STC unit, which was available in a modular form. This permitted alterations to the circuitry to be made at will as the components were easy to reach.

Firstly, the 50μsec. de-emphasis section had to be removed, since as a 50μsec. emphasis was not included in the transmitter, the de-emphasis acts effectively as a 3 KHz. low pass filter.

With the removal of the de-emphasis, the effective bandwidth of the telemetry system as a whole was 50 KHz. This limitation is due to the transmitter section.

The block diagram of the receiver appears in Fig. 6.13.

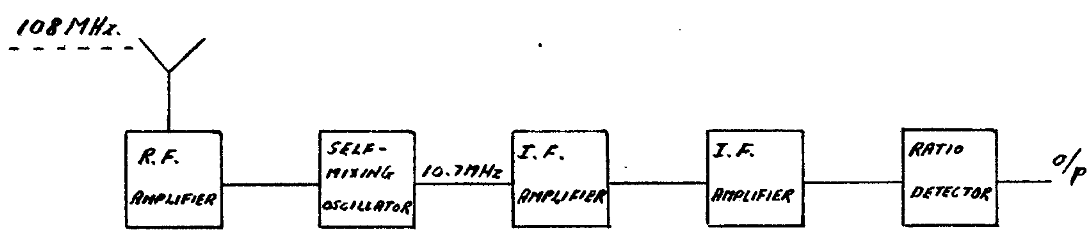


Fig. 6.13

A dipole aerial was constructed using 300 Ω tape to give good matching with the receiver input. This aerial picks up the transmitted signal which is then amplified by the radio frequency amplifier. This amplified signal is then passed through a self-mixing oscillator where a difference frequency of 10.7 MHz. is produced, which is easier to power amplify. This signal is then passed through a two-stage intermediate frequency amplifier where power amplification is effected. These amplifiers have a bandwidth of 200 KHz. (the limiting band-width in the receiver.) The output is finally passed through a ratio detector where the frequency modulation is converted back into an amplitude variation.

The systems specifications are given as follows:

Overall supply voltage: 25 volts D.C.

Circuit stabilised voltage: 5.6 volts D.C.

Current consumption: 7 mA.

Frequency range: 86.5 - 108.5 MHz.

At a signal input level of 2μV., the signal to noise ratio is $\geq$ 12 db.

The detailed design of the receiver is attached as Appendix B.

The output of the receiver is approximately 50 mV. and a SN 72741 op-amp was used to amplify this to a level of 2V. in order to overcome the forward voltage drop across the detector and still be of sufficient magnitude to act as input to the U.V. recorder. The circuit of the amplifier appears in Fig. 6.14.

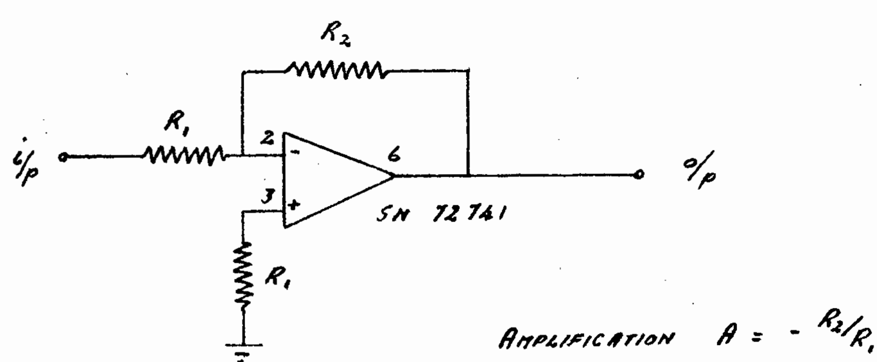


Fig. 6.14

6.2.2 THE FILTERS.

The filters posed quite a problem because of the high rates of cutoff required. It can be seen in Chapter 5 that the required attenuation was at least 25 db. within 30% - 40% of the cutoff frequency and a steady holding of the attenuation below 25 - 30 db.

6.2.2.1 ACTIVE FILTERS.

Active filters were tried initially since these avoid the tedium of winding coils and also the inaccuracies due to mutual coupling. An excellent aid to this design was found in an article by M. Bronzite (1970), which contains detailed tables for the design of active filters with both Chebychev and Butterworth responses. Sixth-order active filters were tested, with rippls orders of 0.5 db. in the pass band. The tables predict a 54 db. attenuation in the first octave.

The filters consisted essentially of a low and a high pass filter in series with one another, which gave a bandpass filter. The filters themselves consisted of sub-sections, the designs of which follow below.

a) Low pass filters.

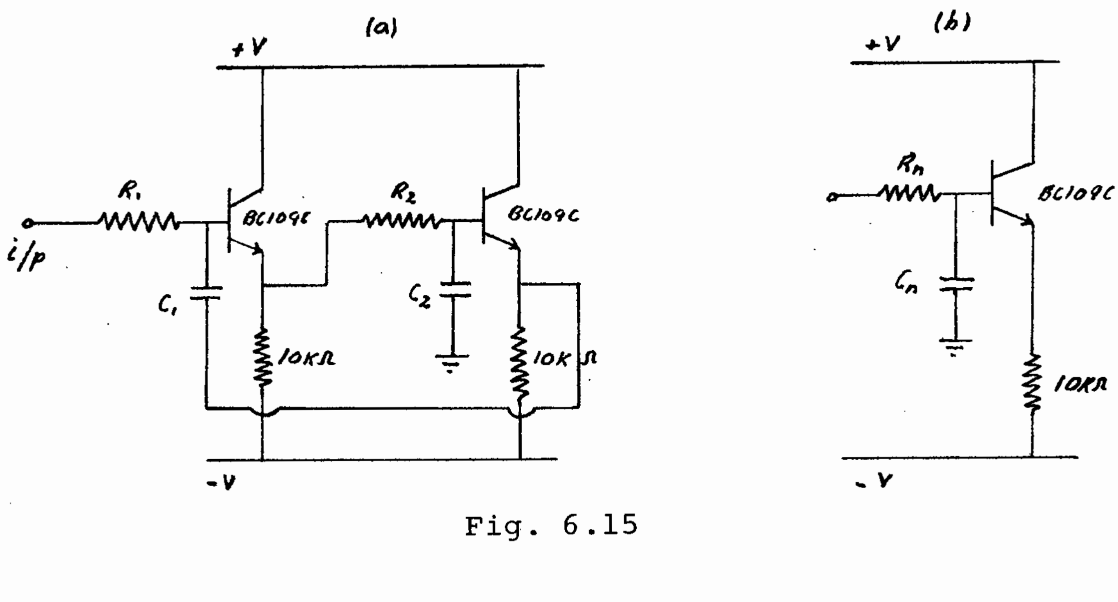


Fig. 6.15

$R_nC_n = \frac{T_n}{f_m}$ (n indicates the position of the RC sub-section in the cascaded filter.)

where $f_m = \frac{f_{3db}}{\beta}$ f_{3db} is the frequency at which cutoff is desired.

β is given in a table in the article and is dependent on the ripple order (m) and the order of the filter (N).

T_n is also given in a table in the above article.

The sub-sections are then cascaded to form the required order of filter. If the filter order is an odd number, the circuit in Fig. 6.15 (b) is added as the last stage. The tables from the article appear in Appendix C.

b) High pass filter.

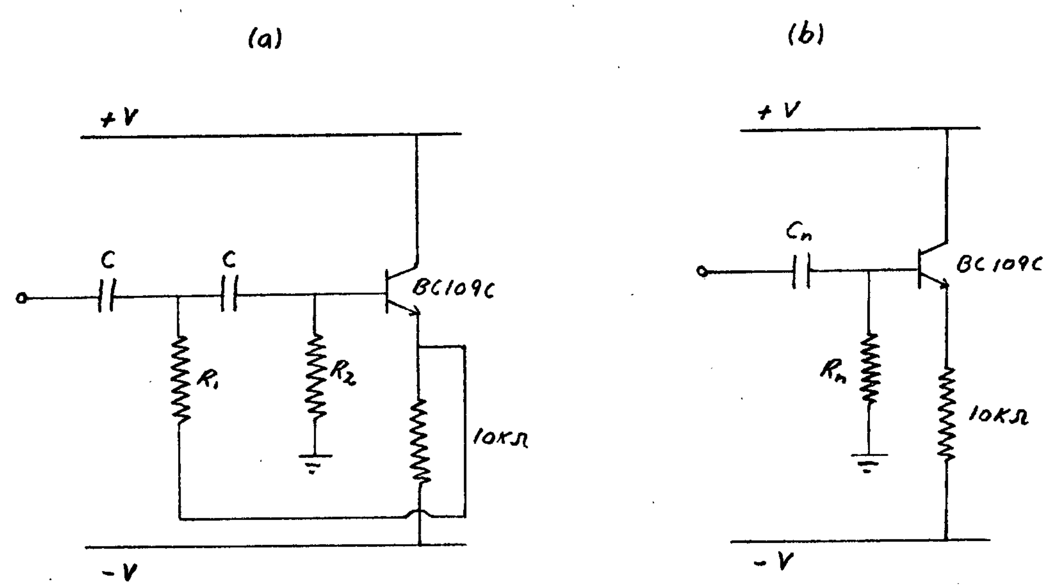


Fig. 6.16

For (a):-

Now $R_1 = \dfrac{T_1}{2Cf_m}$

and $R_2 = \dfrac{2T_2}{Cf_m}$

Here however, $f_m = f_{3db} \times \beta$. f_{3db} and β are the same as for low pass filters. T_n is found in a separate table also in Appendix C.

Fig. 6.16 (b) is again used as the last stage if the order of the filter is an odd number.

The cutoff of these filters was, however, found to be insufficiently sharp for the separation of the frequencies in the system, which were very closely spaced due to the limited available bandwidth. Increasing the filter order was also found to be impractical. Due to the 54 db./octave attenuation, the absolute minimum spacing between channels was half an octave.

6.2.2.2 PASSIVE FILTERS.

Passive M-derived filters have the advantage over the above active filters in that the sharpness of cutoff could be determined almost independently of the filter order. Although very sharp cutoffs limited the overall attenuation, the addition of the prototype section overcame this difficulty. Consequently, the desired attenuation could be achieved within 5% of the cutoff frequency; something well out of the scope of the active filter design.

The problem of accurately wound inductances was overcome by using No. LA 4245 ferrite cores. The inductances of these cores could be very finely set by controlling the position of the centre core. (These cores were wound with gauge 36 copper wire.)

As in the case of active filters, low and high pass filters were cascaded in order to give a band-pass filter. The basic filter consisted of four sub-sections.

1) Two matching networks, one at the input and one at the output.
2) An M-derived section.
3) A prototype section.

The matching circuits provided optimal matching characteristics with other circuits and prevented any unbalance in the filter characteristics due to loading.

The M-derived section provides the very sharp cutoff required as in Fig. 6.17. The attenuation, however, immediately decreases and it is to compensate for this effect that the prototype sections were added.

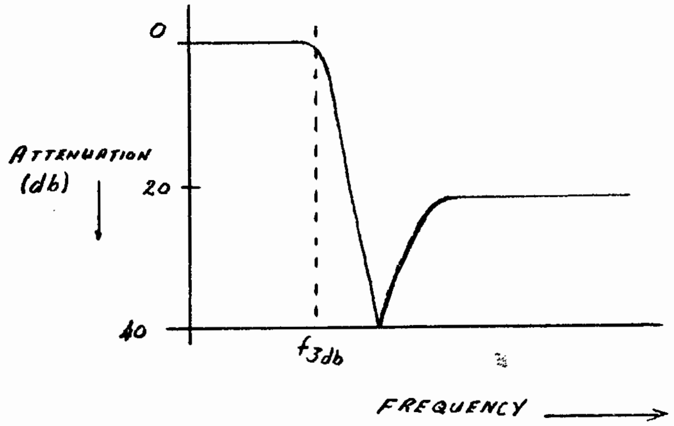


Fig. 6.17

a) Low pass filters.

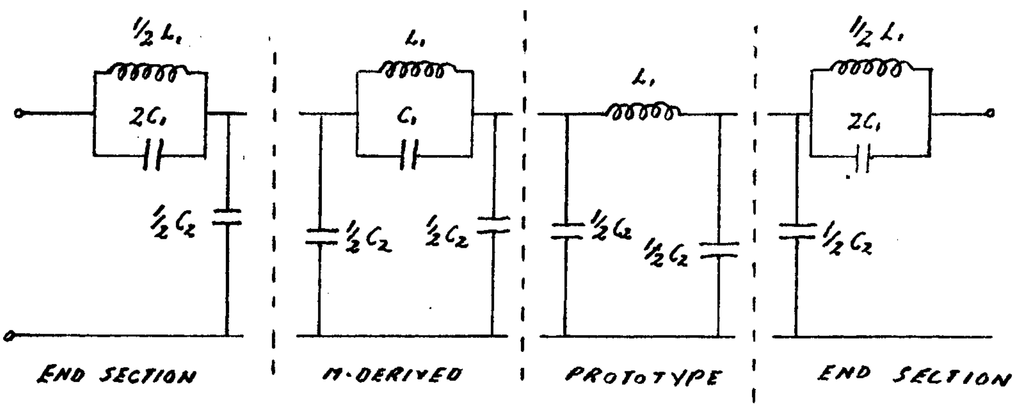


Fig. 6.18

The initial calculation is the determination of m which is to be used in determining all the other circuit values. M is a measure of the sharpness of cutoff and is different for each sub-section.

$m = \sqrt{(1 - (f_2/f_\infty)^2)}$ where f_2 is the frequency at which cutoff begins and f_∞ is the frequency at which maximum attenuation is desired.

The values of m used for the different sections were:-

1) End sections: $m = 0.6$
2) M-derived section: $m = 0.3$
3) Prototype: $m = 1.0$

The second calculation to be made is:

$$L_k = \frac{R}{\Pi f_2}$$

where R is the load resistance and f_2 is as above.

$$C_k = \frac{1}{\Pi f_2 R}$$

The determination of the actual circuit values then follows:

$$L_1 = mL_k$$

$$C_1 = \frac{1 - m^2 C_k}{4m}$$

$$C_2 = mC_k$$

b) High pass filters.

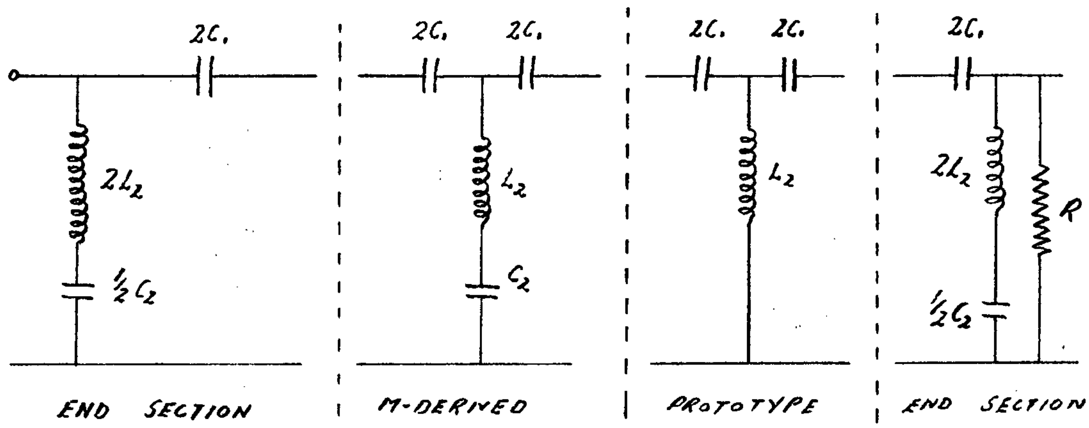


Fig. 6.19

Here $m = \sqrt{1 - (f_\infty/f_1)}$ where f_1 is the cutoff frequency and f_∞ is the frequency at which maximum cutoff is desired.

The values of m used were the same as those for the low pass filter sections.

Also:

$$L_k = \frac{R}{4\Pi f_1}$$

$$C_k = \frac{1}{4\Pi f_1 R}$$

R is the load resistance and f_1 is as above.

The determination of the actual circuit values then follows:

$$C_1 = \frac{C_k}{m}$$

$$C_2 = \frac{4m}{1-m^2}C_k$$

$$L_2 = \frac{L_k}{m}$$

The desired frequency bands are:
0 - 2 KHz., 7 - 12 KHz., 17 - 21 KHz., 29 - 33 KHz., and 45 - 50 KHz., and the calculated component values for these required filters appears in Appendix D. Included in Appendix D is also the final filter characteristics, showing how close they are to the desired ones.

6.2.3 AMPLITUDE DEMODULATORS.

For amplitude demodulation, simple detectors were required to filter out all frequencies above 2 KHz., the maximum e.m.g. frequency of interest. The detectors had to sufficiently attenuate the carrier signals to make their effect unnoticeable (less than 5%) on the output e.m.g. Consequently, a simple diode detector was used. The diode was used to cut off the negative half of the amplitude modulated signal.

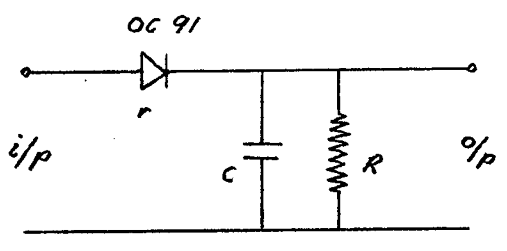


Fig. 6.20

The value of C must be such that it will follow the variations of the modulating signal, i.e., the time constant of the charging circuit (r and C) must be less than or equal to the minimum time constant of the e.m.g. signal. (The minimum time constant is equal to the reciprocal of the maximum frequency.)

$$rC \leq \frac{1}{f} = T \qquad \text{where } r = 500\Omega$$

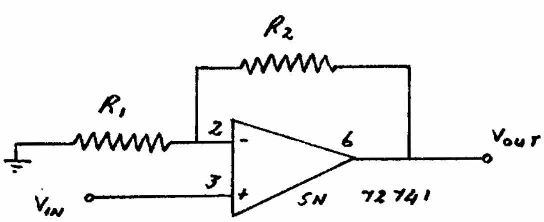
$$\text{and} \quad f = 2 \text{ KHz.}$$

$$\therefore \; C = 1\mu F$$

and R must be such that the capacitor C, in discharging, will follow the 2 KHz. modulating signal.

$$\therefore \; R = r = 500\Omega$$

Since the output of this follower is needed to drive a U.V. recorder or any other type of recording apparatus, a voltage follower (Fig. 6.21) was placed at the output, as otherwise loading of the diode follower would distort the output waveform.



$$V_{OUT} = \left(\frac{R_1 + R_2}{R_1}\right) V_{IN}$$

Fig. 6.21

Letting $R_2 = 1K\Omega$ and $R_1 = 10K\Omega$

$$V_{out} = 1.1 \, V_{in}$$

6.2.4 FOOTSWITCH DECODER.

6.2.4.1 FILTERS.

On being received and de-multiplexed, the two footswitch frequencies were separated out, with second order active filters which were excellent for this application since the frequencies 400 Hz. and 2 KHz. are sufficiently far apart.

a) To pass the 400 Hz. signal and attenuate the 2 KHz. one, a low pass filter with a 3 db. cut-off at 500 Hz. was used.

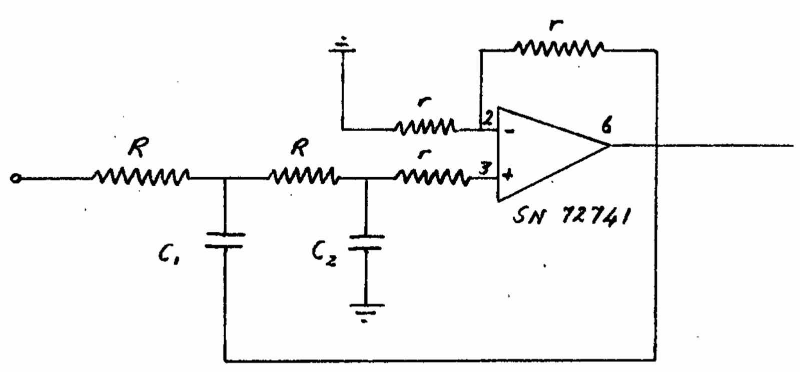


Fig. 6.22

From the article mentioned in 6.2.3.1

$$C_1 = \frac{2T_1}{Rf_m} \qquad \text{and } C_2 = \frac{T_2}{2Rf_m}$$

and f_m = 500 Hz. T_1 = 0.24679 T_2 = 0.14498

Letting R = 10KΩ

C_1 = .1μF

and C_2 = .01μF and r = 10KΩ

b) To pass the 2 KHz. signal and attenuate the 400 Hz. one, a high pass filter with a 3 db.

cutoff at 2 KHz. was used

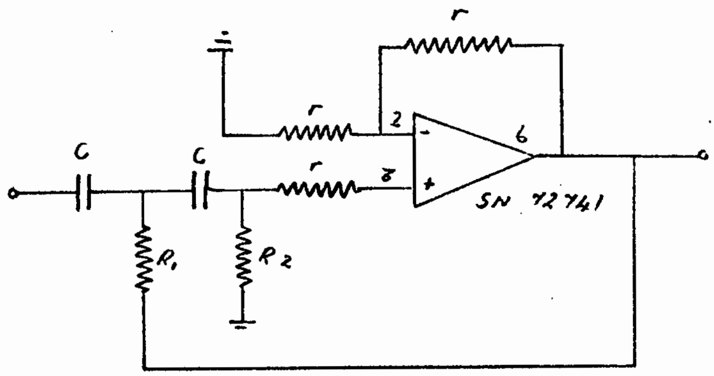


Fig. 6.23

Once again referring to the above article by Bronzite (1970):

$R_1 = \frac{T_1}{2Cf_m}$ and $R_2 = \frac{2T_2}{Cf_m}$

f_m = 2 KHz. T_1 = 0.10264 T_2 = 0.17472

Letting C = 0.01μF

R_1 = 2.56KΩ

R_2 = 17.5KΩ

r was made equal to 10KΩ

6.2.4.2 THE FOLLOWERS.

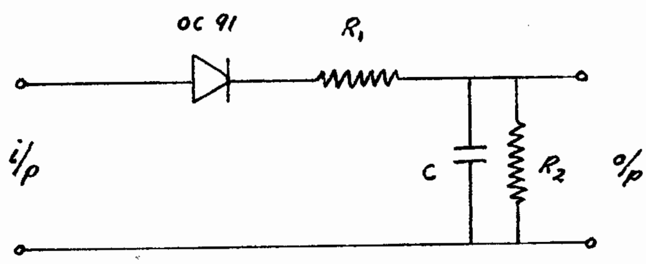


Fig. 6.24

Once again, an ordinary diode follower was used. The diode half-wave rectifies the sine wave input. The time constant of the charging circuit (i.e. R_1 and C) must be less than or equal to the time period of the charging waveform.

$$\therefore \quad R_1C \leq \frac{1}{400}$$

$$\text{Let} \quad R_1C = \frac{1}{4000}$$

$$\text{and if} \quad R_1 = 1K\Omega$$

$$C = .25\mu F$$

Similarly, the time constant of the discharging circuit (i.e. R_2 and C) must be greater than or equal to the period of the charging waveform and yet much less than the actual period of walking.

$$\therefore R_2C \geq \frac{1}{400}$$

$$= \frac{1}{50}$$

This satisfies the second condition as 50 Hz. is much faster than the step frequency.

$$R_2 = 80K\Omega$$

A similar calculation may be performed for the 2 KHz. signal.

6.2.4.3 THE COMPARATOR.

In order to create a square pulse from the output of the follower, a voltage comparator was used. This comparator switches between two output voltages, -0.5 and 2.5 volts as the input passes the 100 mV. level. When the input is below 100 mV.

the output is 2.5 volts and when above, the output is -0.5 volts. An LM 710 voltage comparator was used.

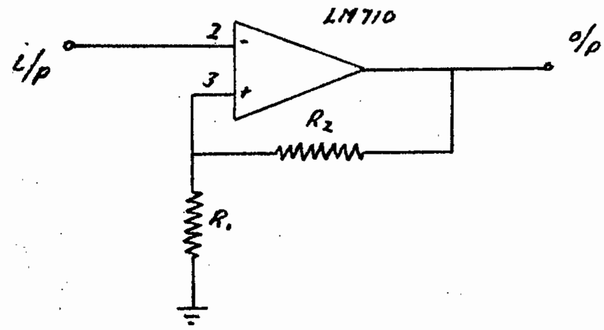


Fig. 6.25

The ratio of R_2/R_1 was very large.

Actual values used were:

R_1 = 1KΩ

R_2 = 1MΩ

6.2.4.4 THE OUTPUT SUMMER.

The idea here is to create three voltage levels indicating:

a) Toe signal present.

b) Heel signal present.

c) Both present.

For this purpose a summing amplifier was used. The output of the "toe" comparator was halved and that of the "heel" comparator multiplied by one.

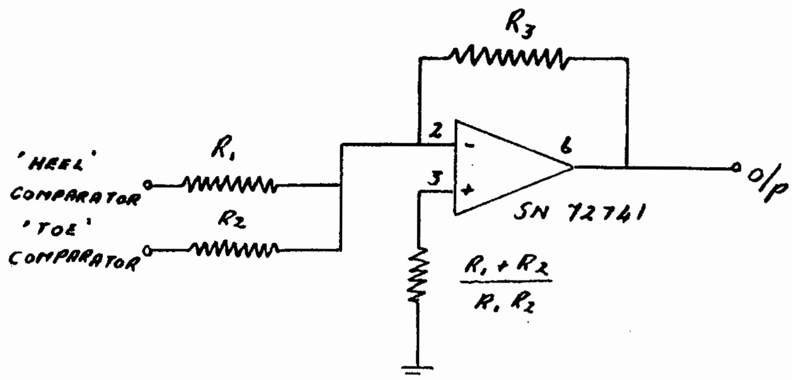


Fig. 6.26

$$\frac{R_3}{R_2} = \frac{1}{2}$$

and $$\frac{R_3}{R_1} = 1$$

$$\therefore R_1 = R_3 = \tfrac{1}{2}R_2 = 1K\Omega$$

6.2.5 RECORDING.

The recording of the e.m.g.s and the footswithes was done on an Ultra-Violet recorder. This instrument was a 12 channel pencil galvanometer recorder from S.E. Laboratories. SMI/L galvanometers were used which had a frequency response of 1,600 Hz. The recording was done on Oscilloscript D, Agfa U.V. paper.

CHAPTER 7.

EVALUATION OF THE SYSTEM CAPABILITIES.

7.1 INTRODUCTION.

In evaluating any system, there are certain basic criteria which should be investigated. Firstly, it is important to note the degree to which the initial specifications have been met in the final system. It is also of value to draw a comparison between the system designed and another existing system, and then from this to consider whether some changes in the initial specifications might be suggested. In addition, possible improvements in the design and alternative methods of construction should be considered before summing up the scope offered by the new system.

7.2 SPECIFICATIONS.

All the initial specifications laid out in Chapter 5 were fulfilled. These included the channel specifications for the four e.m.g. channels and the one foot-switch channel, a total weight of about 750 gm. carried by the subject as well as a size within the limits laid down. Several subjects, on whom the apparatus was tested, all experienced no great discomfort, nor felt unduly restricted in normal walking. It was felt, however, that in the case of more strenuous measurements such as running or very active sport, that smaller amplifiers and modulators ought to be considered.

The initial range specification of 50 metres in open air was also fulfilled, although care had to be taken against the presence of large metal apparatus in the vicinity, as these often created interference.

7.3 COMPARISON WITH DIRECT LINE SYSTEM.

Recordings were made with several different subjects walking normally in a straight line. For the purposes of these tests, only three e.m.g. channels and the one footswitch channel were used. The use of three instead of four e.m.g. channels was due to limitations in the recording apparatus available and not to the telemetry system itself.

The three muscles monitored were tibialis anterior, gastrocnemius, and the quadriceps group, all of the left leg, while the footswitch channel was used to monitor the foot ground contact pattern of the right foot.

In Fig. 7.1, the recordings which were collected using the telemetry system appear, whereas in Fig. 7.2, the recordings are those from the direct line system at Princess Alice Hospital in Retreat.

A slight noise effect is noticeable in the footswitch of the telemetry system. This was due to a slight inefficiency in the footswitch. A far stricter tolerance and perhaps a different design might be necessary for this ffotswitch. As can be seen,

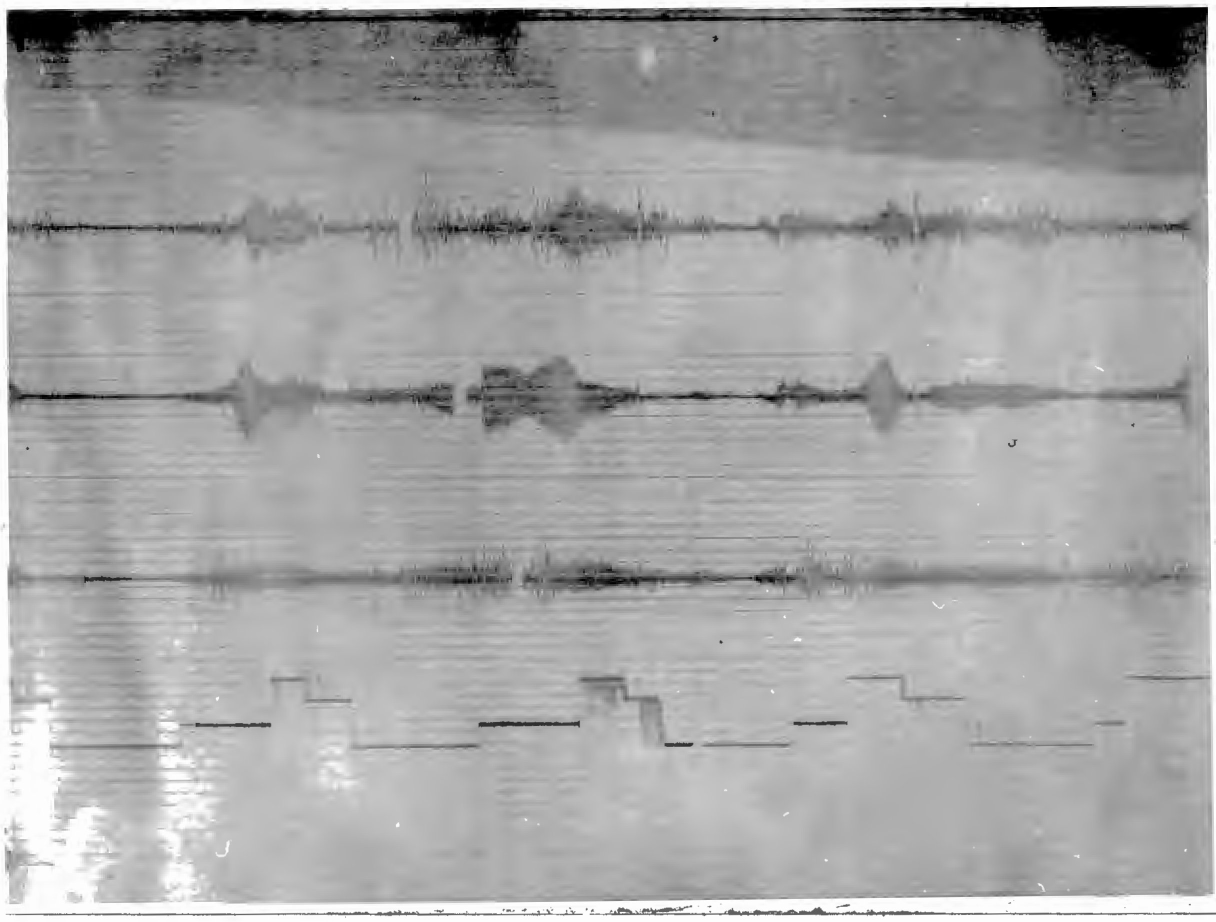


Fig. 7.1

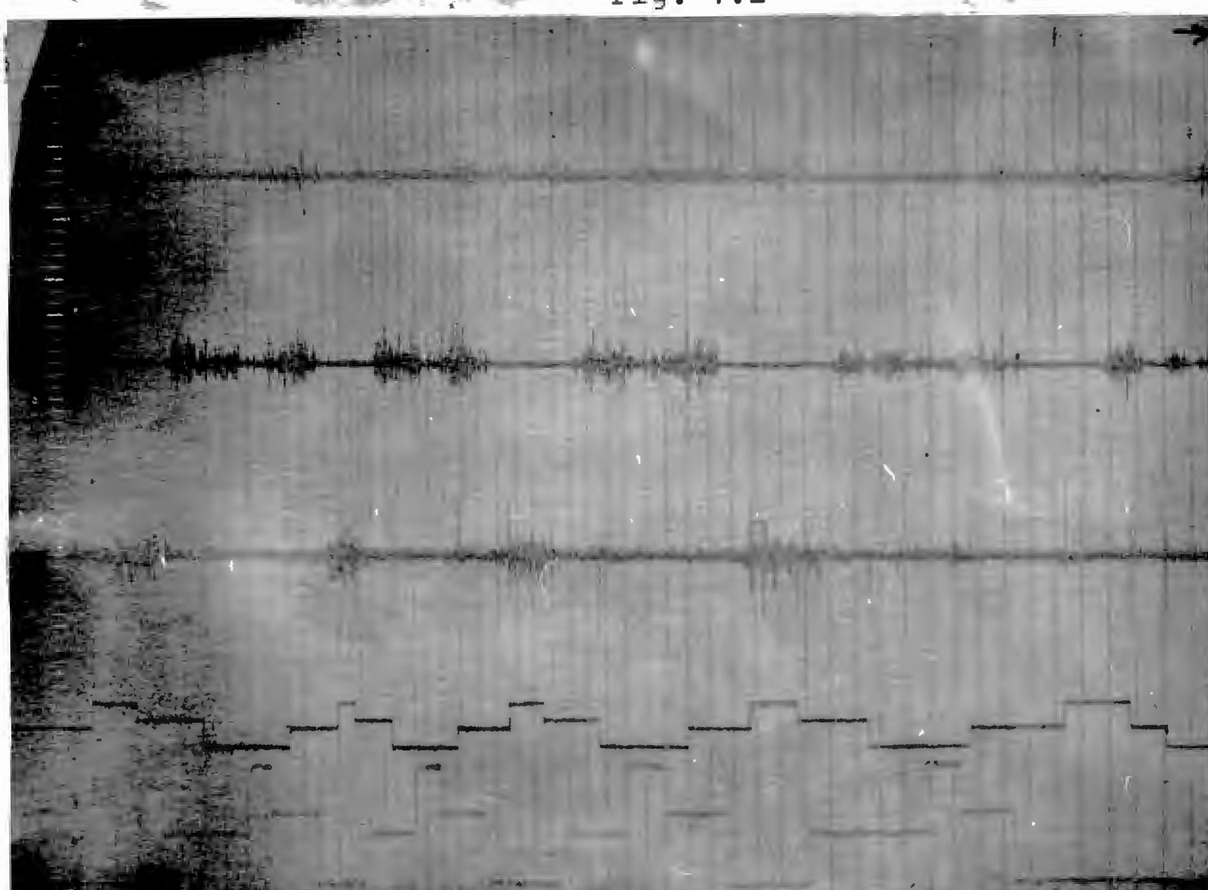
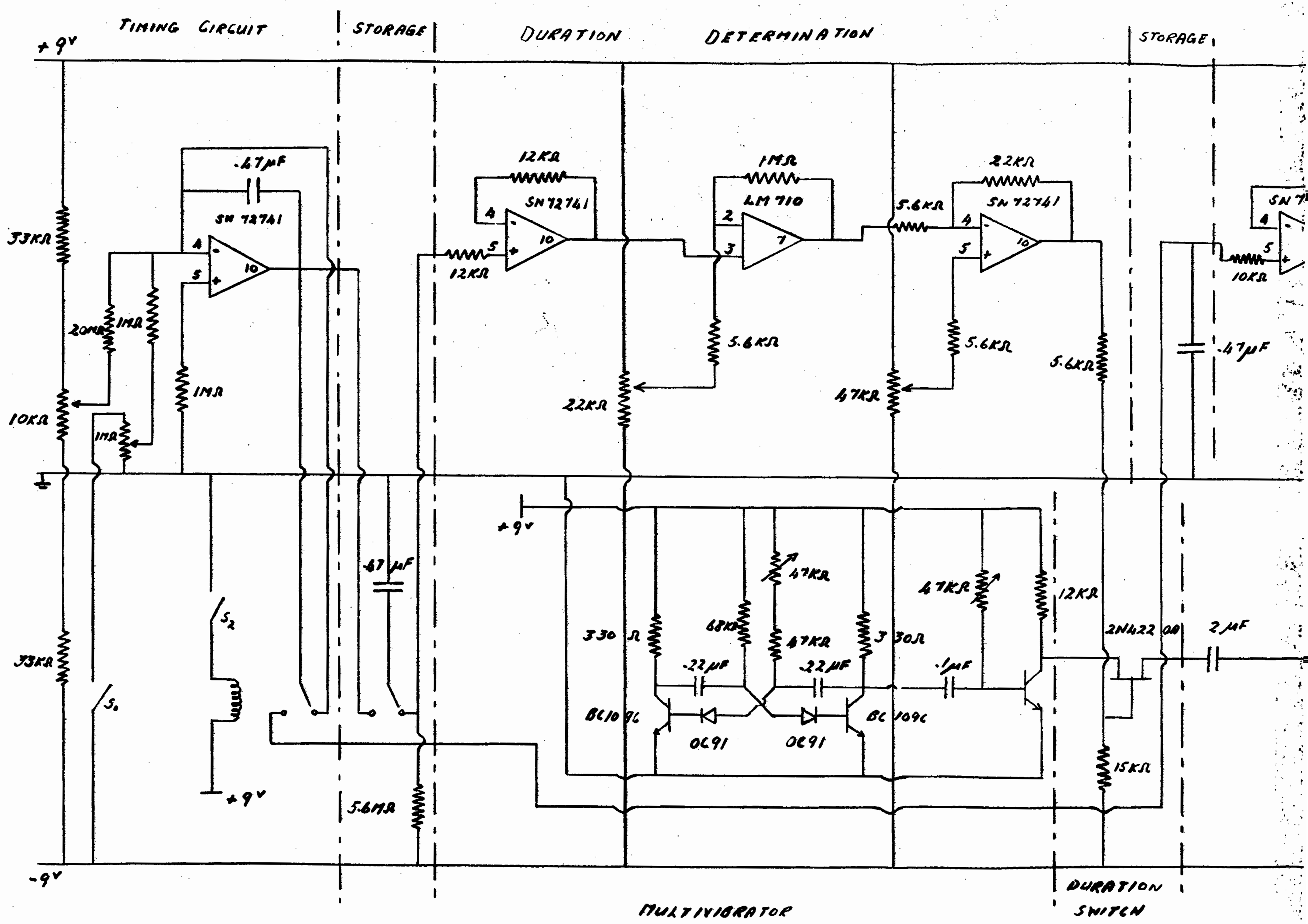
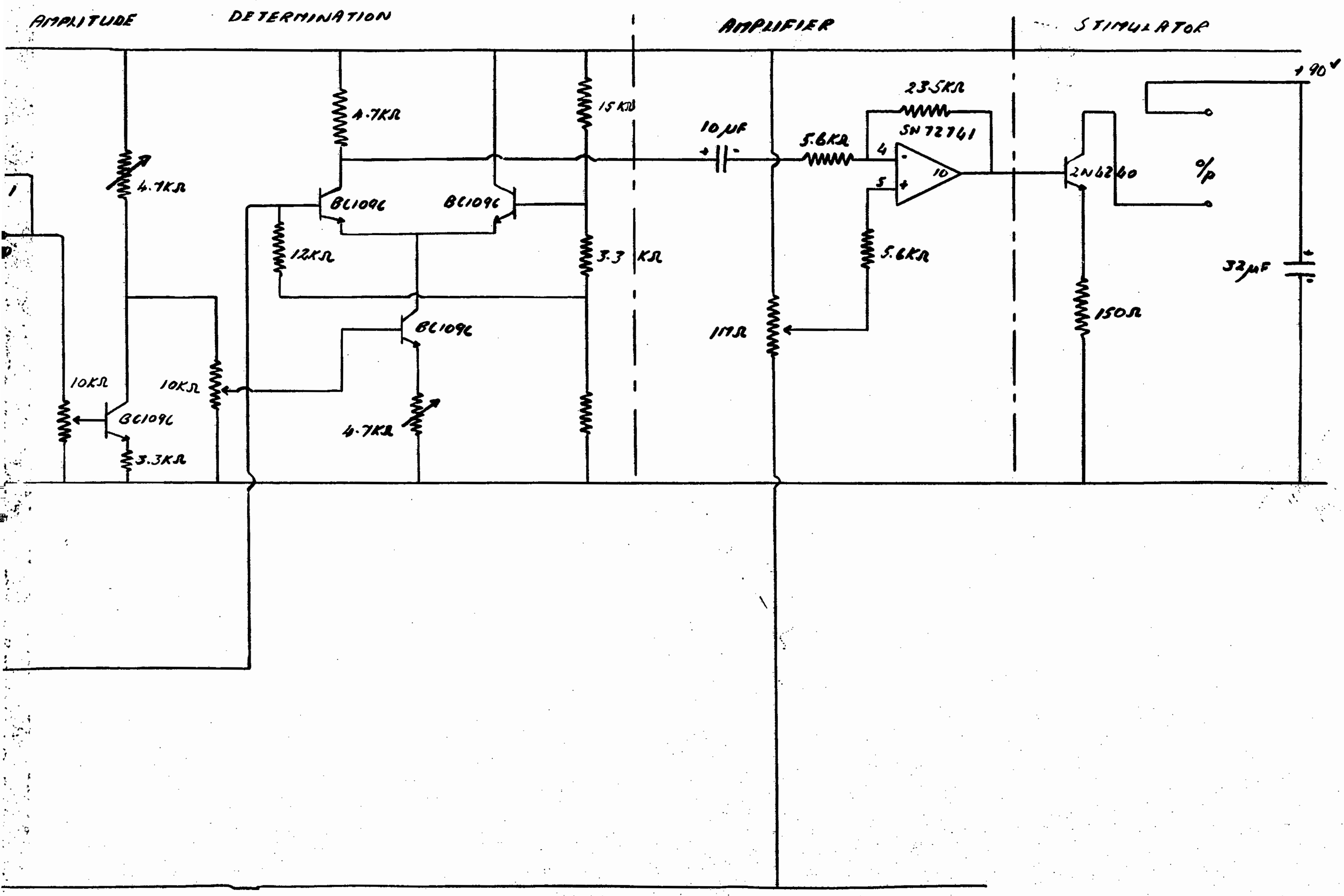


Fig. 7.2



TIMING CIRCUIT
STORAGE
DURATION
DETERMINATION
STORAGE
+9V
-9V
33KΩ
10KΩ
20MΩ
1MΩ
1MΩ
1MΩ
.47μF
SN 72741
12KΩ
12KΩ
SN 72741
22KΩ
1MΩ
LM 710
5.6KΩ
5.6KΩ
47KΩ
2.2KΩ
SN 72741
5.6KΩ
5.6KΩ
10KΩ
.47μF
.47μF
5.6MΩ
S2
S0
+9V
+9V
330Ω
68KΩ
47KΩ
47KΩ
.22μF
.22μF
BC109C
OC91
OC91
BC109C
330Ω
47KΩ
12KΩ
.1μF
2N422
OA
2μF
15KΩ
MULTIVIBRATOR
DURATION SWITCH



AMPLITUDE
DETERMINATION
AMPLIFIER
STIMULATOR
+90V
4.7KΩ
4.7KΩ
15KΩ
10μF
5.6KΩ
23.5KΩ
SN72741
4
5
10
2N6260
o/p
BC109C
BC109C
12KΩ
3.3KΩ
5.6KΩ
32μF
BC109C
1MΩ
150Ω
10KΩ
10KΩ
BC109C
4.7KΩ
3.3KΩ

besides this interference the system's results compare favourably with those of the simpler direct line system.

7.4 ALTERNATIVE SUGGESTIONS.

Perhaps the most important improvement could be made in miniaturising further the circuits carried by the patients. This would have two definite advantages; firstly, in the greater freedom of movement afforded to the subject, since smaller circuits would be easier to attach inconspicuously to the patient; and secondly, since they could be attached more firmly, they would thus assist in the elimination of any movement artefact. Miniaturisation could be accomplished by thick film techniques to construct the circuits themselves. Enclosing all the circuits in epoxy instead of the perspex boxes used would further decrease the size.

It might be worthwhile to test an FM/FM version of the system as opposed to the AM/FM version used. This might offer a slightly improved result, by serving to decrease the noise.

As mentioned in 7.2, it might be advantageous to increase the range specification slightly; 100 metres would be more than sufficient to cover the wider range of activities envisaged.

7.5 CONCLUSION.

The system which was designed and built, successfully fulfilled the initial specifications decided upon. Though perhaps not good enough for vigorous sports, the system would fare very well in the monitoring of gait patterns in a normal environment and during non-strenuous sports. The system adequately answers the need for multi-channel e.m.g. transmission, a very valuable tool for the comprehensive classification of walking patterns which is so important to paralysed patients in various classes and amputees alike.

CHAPTER 8.

REVIEW OF THE LITERATURE OF MUSCLE STIMULATION.

8.1 INTRODUCTION.

When it was realised that muscle contractions were effected by the transmission of a type of electrical signal, along the motor nerves to the muscle fibres, the question inevitably arose whether it might not be possible to contract a muscle by simply applying an external electrical signal to it. Muscular contraction, due to external electrical impulses, had in fact been discovered in the early eighteenth century, though the processes involved were not fully understood at the time and consequently, incorrect conclusions were made.

Luigi Galvani's (1737 - 1798) experiments with frogs' legs were among the first to illustrate the electrical stimulation of muscles. In the experiment he connected the muscle of the leg to an iron grating and the nerve to a length of copper wire which formed a closed circuit with the grating. He never realised, however, that the resulting contraction was due to the voltage created across the junction of two dissimilar metals, but felt this result somehow revealed the presence of a new source of electricity in the animal body.

The first person to introduce wide usage of electrical stimulation was Duchenne de Boulogne in

1867. At that time, the main application of muscle stimulation was in electrodiagnosis.

8.2 USES OF MUSCLE STIMULATION.

The idea of using electrical muscle stimulation as a functional aid for paralysed patients is a rather new innovation, but electrical stimulation has been used extensively for quite some time in diagnosis and in the treatment of muscular and nervous disorders.

8.2.1 ELECTRODIAGNOSIS.

(Joseph Goodgold, Arthur Ebestein, 1972)

With the use of muscle stimulation, certain muscular diseases or injuries are easier to diagnose than with straightforward e.m.g. measurements. The results obtained with stimulation are also less dependent on patient co-operation and hence they offer a more reliable indication of the patient's condition.

Percutaneous or external nerve stimulation is a very simple diagnostic tool to use and yet it may reveal significant information as to the state of the peripheral neuromuscular system. Electrical stimulation of the facial muscles provides an excellent method for the evaluation of facial paralysis.

Buchtal, F. and Rosenfalck, A., (1966) have done a lot of work towards improving techniques for nerve conduction velocity experiments. This new diagnostic aid, though not fully developed, has already achieved quite significant results.

A very common method for determining muscle excitability is the strength-duration curve. In this test, a graph is drawn in which the stimulus amplitude required to give a minimum muscle contraction is compared to the pulse duration of the stimulus. This curve varies as the nerve regenerates and so can be used as an indication of this regeneration.

A book by Joseph Goodgold and Arthur Ebestein gives an excellent summary of the latest developments in electrodiagnosis.

8.2.2 ELECTROTHERAPY.

With no activity, a muscle generally loses its viability and the cells tend to die off. For this reason it is essential to keep up a certain amount of muscle activity in those situations where voluntary contractions are difficult or impossible to achieve. Electrical stimulation of the muscles can provide the necessary amount of involuntary exercise to keep up the required muscle tone.

This is especially important in the case of denervated muscle fibres where it is important to keep the muscles in a state of vitality till nerve regeneration occurs. A significant example of this type is in the case of facial palsy.

Electrical stimulation is, however, not the only method of retaining muscle tone and many physiotherapists feel that passive movements and massage are equally good.

Muscle stimulation also helps to build up weak muscles by exercise, and in the case of people long incapacitated, to re-educate them in the use of the affected muscle. An interesting case of this type is with quadriplegic patients who may require the use of splints for effective hand control. Good wrist extensors are required for control of these splints and it is to strengthen and re-educate these muscles that electrical stimulation is used. (C. Ruch.)

8.3.1 PHYSIOLOGICAL BACKGROUND TO FUNCTIONAL ELECTRICAL STIMULATION. (Dmitrijevic, M.R. et al; 1968)

The neural interconnections of the motor nervous system are innumerable, and interwoven with a large degree of complexity. To simplify the understanding of this system, it has been divided into upper and lower motoneurones. The upper motorneurone is defined as having its nucleus in the cerebral cortex, while its fibre extends down along the spinal cord. Here it connects with the cell body of the lower motorneurone before its fibre leaves the cord and threads its way through the body to the muscle it controls.

The lower motorneurones consist, however, of both alpha and gamma motorneurones. Alpha motorneurones have large cells and thick axons and they activate the skeletal muscle itself. Whereas the gamma motorneurones are smaller and they ennervate the muscle spindles, the small fibres responsible for muscle tone. The spindles also relay information about muscle tone to the alpha motorneurones.

The alpha lower motorneurones thus have direct control over muscle contraction but they are activated and controlled in turn by the upper motorneurones, the gamma lower motorneurones, and variousother sensory nerve paths. There exists thus, two regions of control of the motor nervous system; the cerebral cortex and the localised reflex loops. The cerebral cortex may be regarded as the gross controller while the reflex arcs control the complicated process of automatic regulation.

8.3.2 MOTORNEURONE LESIONS.

Motorneurone lesions are of two main types: upper and lower. Upper motorneurone lesions leave the lower reflex loops intact and excitable. Patients with this type of lesion exhibit little or no voluntary control of the affected muscle and spasms or spontaneous involuntary contractions are very common. It is this type of person, however, who may benefit from functional electrical stimulation, as a means still exists of exciting the muscles, viz., via the lower motorneurones. On the other hand, lower motorneurone lesions are unsuitable for this type of functional aid as the nerves which excite the muscles have been destroyed.

8.3.3 AFFERENT AND EFFERENT STIMULATION.

Efferent stimulation is the stimulation of the peripheral motor nerves themselves to effect muscular contraction. This method was introduced by Liberson et al (1960) and quite a lot of developed of this work has been done by Vodovnik et al (1966) and Milner, Quanbury and Basmajian (1970). Efferent stimulation is the most widely used and will be discussed below. Afferent stimulation which is the stimulation of the sensory nerves and which thereby activates the localised reflex arcs, has been experimented with by Dimitrijevic et al (1968). This has definite advantages over the former method especially in so far as pain, fatigue and inhibitory control of muscular contractions are concerned. Certain problems, however, still have to be overcome before this method can be used effectively.

8.4 STIMULATION PARAMETERS.

The ideal situation would be to have a stimulus waveform the same size and shape as that of the nerve action potential, and for this stimulus to be distributed in time and position, in much the same way that the reflex loops of the lower motorneurones accomplish this load sharing among the muscle fibres. This, however, is not practically feasible, especially with efferent stimulation and consequently a lot of experimental work has been done to achieve the optimum results with practical stimulations.

8.4.1 STIMULATOR FREQUENCY.

According to Chandler and Sedgewich (1971), the normal motor unit is activated about thirty times a second and they feel that stimulating artificially at higher frequencies is one of the reasons for early fatigue in the stimulated muscle. Liberson et al (1960) used 30Hz. as their stimulator frequency. When using special force measuring apparatus, there seems to be a maximum force limit at just over 50Hz., after which the force effected by an artificial contraction begins to decrease. The value of 50Hz. has been found by independent researchers Milner et al (1970) and Vodovnik et al (1965), and is now widely used for most muscles.

8.4.2 STIMULATOR PULSE SHAPE.

Vodovnik et al (1965) investigated several types of stimulus waveforms.

1. A.C. waveforms: a) Sine waves.
   b) Alternating pulses.
2. D.C. waveforms: a) Rectangular pulses.
   b) Exponential pulses.

Though the pain caused by any specific waveform was no greater than that of any other, they found that the d.c. signals required about one percent of the power that the a.c. signals required, to effect the same contraction. Almost all experimenters in this field now use trains of d.c. rectangular pulses to effect contraction.

8.4.3 PULSE WIDTHS AND TRAIN DURATION.

An extensive study of stimulus pulse widths has been made by Crochetiere et al (1966) in which they constructed curves of stimulus amplitude versus torque with pulse width as a third variable. Liberson et al (1960) felt that there was no specific preference among their subjects for a definite pulse width in the range 20 to 300µsec. Most experimenters now use a pulse width of about 2 msec.

Train duration is very much dependent on the specific application to which a functional stimulator will be put. It is, however, important to realise that fatigue increases greatly with increased train duration. Kralj et al (1971) have investigated the effect of this variation on muscle fatigue.

8.4.4 STIMULUS AMPLITUDE.

Stimulus amplitude is another parameter which is very much dependent on the specific application of the functional stimulation and also on the muscle being stimulated.

Constant current stimulators are preferred by most researchers and peak currents from 10 to over 100 mA. have been used.

8.4.5 ELECTRODE SIZE AND POSITIONING.

There are two main criteria to be considered for optimal electrode size. These are patient comfort and the effect of electrode size on torque. Crochetiere et al (1966) found optimal electrode size for biceps stimulation to be about one inch in diameter. Milner, Quanbury and Basmajian (1969) in stimulating the muscles of the leg, found subjects preferring electrode areas of at least two square inches. These electrodes were placed over the motor point of the muscle while the positive or indifferent electrode, a large electrode of 12 square inches, was placed over the stimulated muscle.

8.5 PROBLEMS ASSOCIATED WITH MUSCLE STIMULATION.

Since neither the stimulus waveforms nor the method by which the stimuli are applied are not ideal, the resultant contractions are not ideal and problems result.

8.5.1 PAIN.

Since the threshold of the sensory nerves is lower than that of the motor nerves, sensory nerves will be fired by any stimulus that activates motor nerves.

Consequently, when the stimulus rises above a certain level, the pain becomes unbearable. Any undue inconvenience, especially in a functional aid, is liable to cause the rejection of it by the paralysed patient. As mentioned above, Milner et al (1969) have investigated ideal electrode size and positions to minimize pain and it is possible to elicit effective contraction with sufficiently little inconvenience to the patients.

8.5.2 FATIGUE.

Muscles during normal activity take far longer to fatigue than those that are stimulated electrically. This is due primarily to the fact that the normal reflex loops stagger the excitation of the muscle fibres, resting some while others are stimulated. With electrical stimulation, all fibres are excited simultaneously and no chance is given for them to recover. Kralj, Trnkoczy and Vodovnik (1971) have extensively investigated the possible minimisation of fatigue but the problem still remains. A solution suggested by Peckham et al (1970) of sequentially stimulating electrodes might go some way to relieve this problem, but further investigation is still required.

8.6 FUNCTIONAL ELECTRICAL STIMULATION.

As was mentioned in the introduction above, using electrical stimulation as a functional aid for paralysed patients should have seemed obvious since Galvani's frogs' legs experiment. However, it was probably the invention of miniature electronic components, such as the transistor, which, by offering the practical possibility of this type of functional aid, gave just that little extra boost that was needed.

8.6.1 THE PERONEAL BRACE.

Liberson and associates designed and tested the first practical functional electrical stimulator in 1960. It was used to aid hemiplegic patients who suffered from drop foot, a common complaint among these patients, causing them to trip over their own feet. The stimulator was activated by a switch under the affected foot, so that whenever the foot was raised, the tibialis anterior muscle was stimulated, causing the foot to dorsiflex. Liberson's brace was extremely succesful, and amazingly revealed that after a period of using the stimulator, some patients re-acquired the ability to dorsiflex voluntarily. A commercial version of this brace is available from Philips, and has been developed by H.J. van Leeuwen and J. Vredenbregt (1969)

8.6.2 PROGRAMMED FUNCTIONAL STIMULATION.

Owing to the success of the peroneal brace, several researchers have felt that vast possibilities exist for programming stimulators to assist paralysed patients in various phases of their walking cycle.

Extrapolated, this could mean that it might one day be possible to have programmed stimulators assisting paraplegics to walk normally.

Coming back to present reality, though, a walking rate dependent peroneal brace has been developed by Kralj, Reberset and Gracanin (1971). They felt that there was an optimum percentage of the swing phase during which the affected leg should be stimulated. They determined this value to be 70% of the swing phase. Their brace, the PO-8, worked very successfully. but they have as yet, not satisfied themselves as to whether a walking rate peroneal stimulator is really necessary.

A three-channel stimulator has been developed at the University of Lubljana by Kralj et al (1971). This stimulator had three stimulus channels, all of which were triggered by the same signal from a foot-switch. The delays in activation, the magnitude and the duration were independently variable for each channel. The real problem associated with this type of passive programming is the inability of the stimulator to take account of the varying walking conditions and situations which a patient might encounter. This very ambitious attempt, however, still requires a certain amount of further development.

A slightly different approach to multi-channel stimulation has been made by Rapley and Milner (1971). Their idea incorporated a handswitch whereby the order of firing was manually controlled.

8.7 CONCLUSION.

The art of multi-channel programmed stimulation is still very much in its infancy. Tremendous advances have been made in quite a short time, a significant contribution coming from the Department of Biomechanics at the University of Lubljana. More information on gait control mechanisms and perhaps better methods of functional aid evaluation are required before wider use can be made of these stimulators.

CHAPTER 9.

ANALYSIS OF THE STIMULATOR.

9.1 INTRODUCTION.

Innumerable aids are available today for hemiplegic, paraplegic, and quadriplegic patients, such that it is not too difficult to provide a patient with many of the amenities available to normal people. Systems have been designed, whereby with simple switch arrays, activated by short breaths, a telephone may be answered, or the volume on a radio altered. Specialised environments have been designed to provide simpler and safer living conditions for patients confined to wheelchairs. Many patients, however, still regard "all these machines" around them as a constant reminder of their disability, and states of depression are very common. Giving back to these people their ability to walk, even in a limited sense, would be a great improvement over the independence offered by the available aids. It is with this aim in mind that researchers have attempted to design muscle stimulus systems to simulate the actions of the normal musculature system (refer to Chapter 8).

The human nervous and musculature system is extremely complex, and for a muscle stimulus system to emulate it in complexity would require an apparatus so large and cumbersome as to make the whole project not feasible. The initial success, however, of the peroneal brace has shown that a gross model can still have a remarkable effect in helping to restore the activity of a paralysed muscle.

9.2 THE WALKING CYCLE.

The walking cycle may be divided into two segments for the sake of analysis: the swing phase, when the foot is in the air, and the stance phase, when the foot is on the ground.

Muscle contractions during the swing phase are not required to exert any large propulsive forces and as such, the magnitude of the contractions need not be too large nor even very finely controlled. What is important in the swing phase, though, is the time synchronisation of the various contractions in order to get a smooth forward swing of the leg. The stance phase, however, requires both the exertion of large forces to propel the person forward, and fine control of the magnitude and time synchronisation of the contractions. This is rather difficult to achieve because of large response lags between the stimulus and the contraction, the movement of the motor point with muscle contractions, and the production of a non-linear force by the stimulus. (There is a lag of 20 - 50 msec. after the beginning of the stimulus and 100 - 200 msec. after the end. Kralj et al, (1971)). An added disadvantage is fatigue, which becomes more significant at the increased amplitudes required for the stance phase.

It is for the above reasons that most systems till now have concentrated on the swing phase. Kralj et al (1971), however, in developing a multi-channel stimulator for the swing phase, extended the stimulation of the quadriceps into the beginning of the stance phase, where it was used to lock the knee, just as the body weight was brought to bear on it. They reported the successful operation of this stimulation effect in the stance phase.

It was consequently felt that the real limitations of a gross stimulator in the stance phase could only be determined after actual experiment, and it was then decided to design and investigate a stance phase stimulator.

9.3 CHOICE OF MUSCLE TO BE STIMULATED.

The gastrocnemius and the quadriceps muscle groups have very definite functions during the stance phase. The gastrocnemius is largely instrumental in giving the forward thrust by extending the ankle, while the quadriceps muscle group is involved in keeping the knee locked so that an effective forward movement can be achieved. The quadriceps muscle is, however, more difficult to stimulate than the gastrocnemius, the activation of two motor points being required to effect a reasonable contraction. It was consequently decided to concentrate initially on the gastrocnemius stimulator. Once a stimulator of this type has been designed, it should not be difficult to use a similar type for the quadriceps group.

9.4 ANALYSIS OF GASTROCNEMIUS ACTIVITY DURING THE STANCE PHASE.

Milner et al (1970) measured the phasic activity of the gastrocnemius during walking for various speeds. It can be seen from the above data that the gastrocnemius is activated for between 40 - 50% of the heel strike to heel strike period, the actual percentage varying from patient to patient, but remaining fairly constant with speed. The point of the onset of activity was not clearly defined but it seemed to coincide very closely with the moment the toe of the other foot left the ground.

The average amplitude of the gastrocnemius activity increased almost linearly with the speed of walking. This was very similar in the case of the quadriceps.

9.5 CONCLUSION.

Keeping the above discussion in mind, it was decided to build a stimulator for a hemiplegic patient to assist him in getting better push-off during the stance phase. A hemiplegic patient was chosen so as to simplify the evaluation of the stimulator. The stimulator was to be able to cope with varying speeds of walking.

CHAPTER 10.

SYSTEM DESIGN.

10.1 BASIC SYSTEM REQUIREMENTS.

There are three basic degrees of freedom which a stimulus signal has, once the frequency, pulse width, pulse shape and electrode positioning have been set at their optimum levels. These are the stimulus duration, the duration between stimulus trains and the stimulus magnitude. It is the complex interweaving of these three parameters for the various muscles that creates the walking pattern. Any programmed stimulation must therefore have the basic system layout as described in Fig. 10.1.

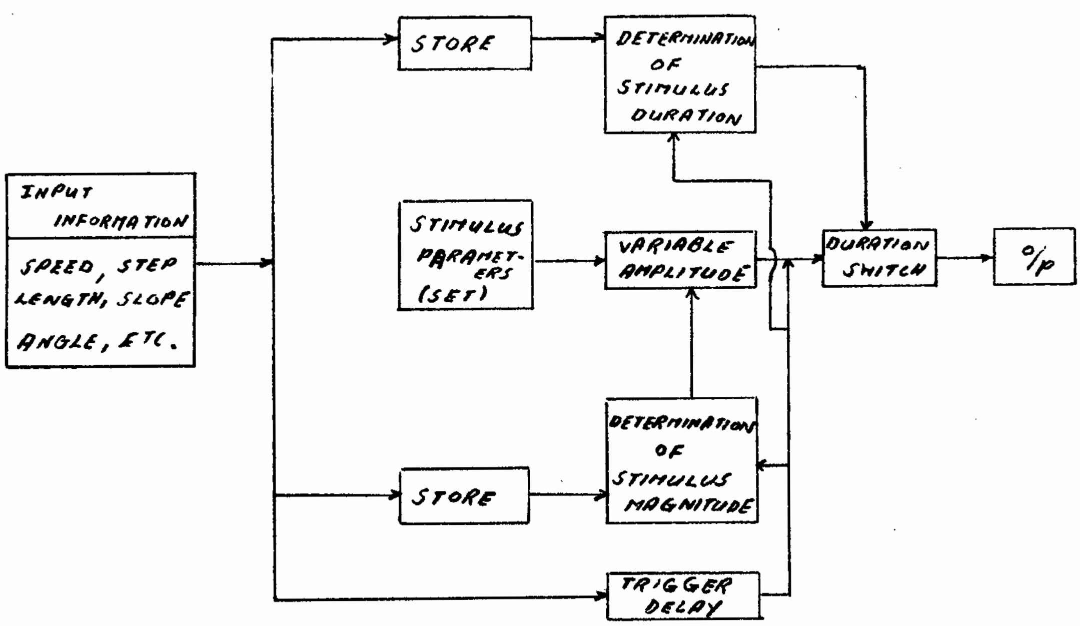


Fig. 10.1

There is initially an input section which collects all the relevant information from the external environment, such as speed of walking, heel strike, etc. The information received here is then stored and used for the determination of the stimulus duration and magnitude. When the trigger is activated, the stimulus magnitude is then automatically set and it passes through to the output for the pre-determined amount of time.

10.2.1 INPUT INFORMATION.

The input information required by a stimulus system, capable of emulating the human locomotor system, would be vast. The system would have to be notified of approaching steps, increasing upward or downward slopes, on corners, to mention just a few parameters. With the system under consideration, however, it was decided to limit its scope to cope only with a variation in the speed of walking.

The initial aim, therefore, was to find some easily measurable parameter which would give an indication of the patients speed, since to actually measure the walking speed directly would be a very complicated affair. An interesting relationship noted by Milner and Quanbury (1970) and depicted in Fig. 10.2 provides a simple answer.

The above relationship was found with no

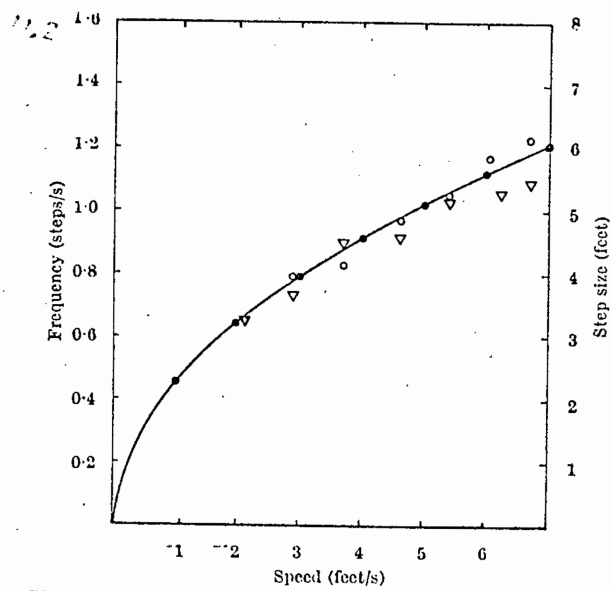


Fig. 3. Step size and step frequency against speed. O, Fitted curve, $y=2{\cdot}28\sqrt{x}$; O, step size; ∇, frequency.

Fig. 10.2

constraints on the subject besides constant speed. It can be seen that the step frequency has a slightly exponential relationship with respect to the speed of walking. For the purposes of this experiment, however, it was felt sufficiently accurate to approximate the relationship to a linear one. This assumption was valid since the idea was to create as simple and straightforward a system as possible, to evaluate its effectiveness, and then to decide whether a more complex version would be necessary. In addition the relationship, with the exception of lower speeds, was sufficiently linear to justify this approximation initially.

10.2.2 MEASUREMENT OF STEP FREQUENCY.

It follows from the above that if the step frequency is known, knowledge of the walking speed would automatically follow. The step frequency is given directly by the inverse of the step period, an easily measureable quantity.

The following methods exist for determining the step period:

a) E.m.g. activity.

b) Foot-ground contact.

Since when measuring the pace period of a subject, it is necessary to have a very clearly defined point at which to begin and end measuring, use of the e.m.g. for this purpose is not advisable. Though e.m.g.s do have a periodic activity, the actual onset of activity is insufficiently defined.

The measurement of foot-ground contact time, however, is a very definite method of determining the step period, there being a sharp transition from one step to the next. This was, consequently, the method used for determining the step period.

Referring to the same paper by Milner and Quanbury (1970a), it could be seen that the toe down period as a percentage of the total step period remained reasonably constant over a range of walking speeds. (This can be seen in Fig. 10.3). Thus by measuring this period, the step period could be directly determined.

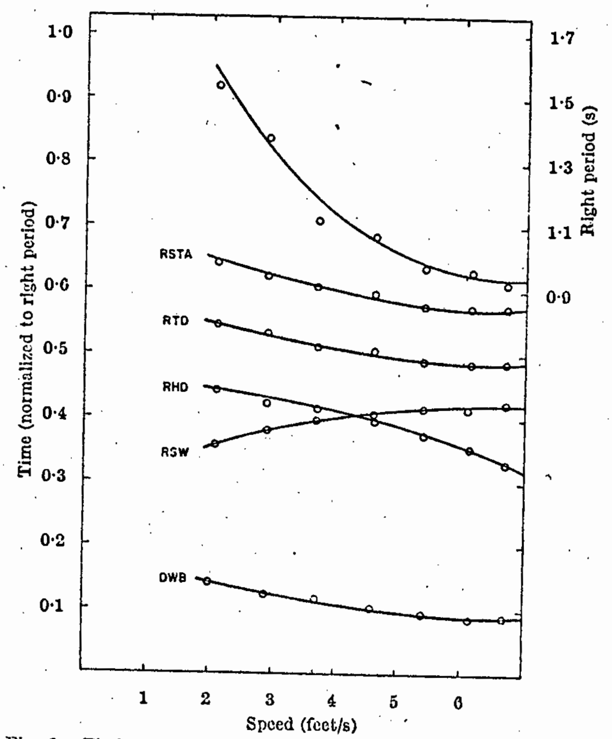


Fig. 2. Timing normalized to right period of footswitch phases for various speeds.

Fig. 10.3

10.3 DETERMINATION OF STIMULATION DURATION.

Referring to the results obtained by Milner and Quanbury (1970b) with respect to the gastrocnemius muscle, it could be seen that the time for which the muscle was active, as a percentage of the step period, varied greatly from person to person (with normal people.) Consequently, the stimulus duration as a percentage of this period must be adjustable for each patient. However, an interesting point to note was that for any single person this percentage remained reasonably constant with varying speeds. The stimulation period could thus be set directly proportional to the step period, measured in 10.2.2 and the constant of proportionality could be the same for all walking speeds. This is illustrated in Fig. 10.4.

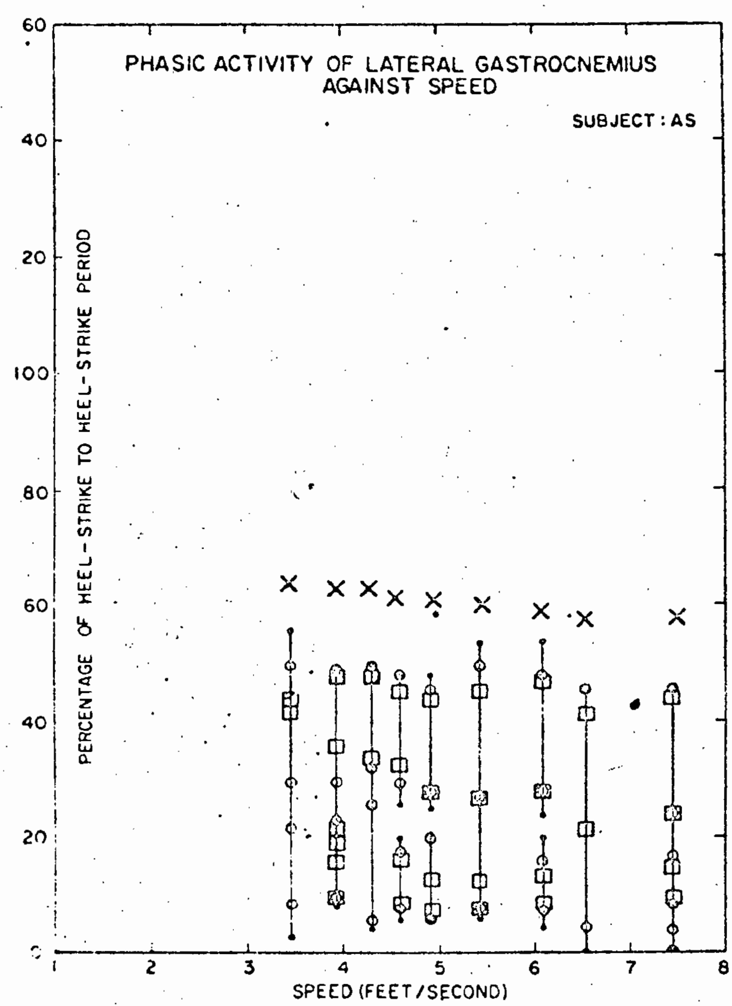


Fig. 10.4

10.4 STIMULUS TRIGGER.

Referring to the graphs in the above paper, e.g. Fig. 10.4, it could be seen that gastrocnemius activity begins just after the heel of the same foot strikes the ground. This coincides fairly closely with the moment the toe of the opposite foot leaves the ground. Consequently, this was initially set as a triggering point for the onset of stimulation, but should this be inadequate, a delay in triggering could always be introduced.

10.5 DETERMINATION OF STIMULUS MAGNITUDE.

Once again referring to the paper by Milner and Quanbury (1970b), it was seen that the average e.m.g. increased almost linearly with speed. For similar reasons to those given in 10.3 it was felt feasible to linearise this relationship for the sake of initial experimentation. Thus as the step duration measured in 10.2.2 increased, indicating a decrease in the speed, the magnitude of the stimulus was made to decrease linearly. The relationship was naturally to be variable, so that it could be adjusted for each patient.

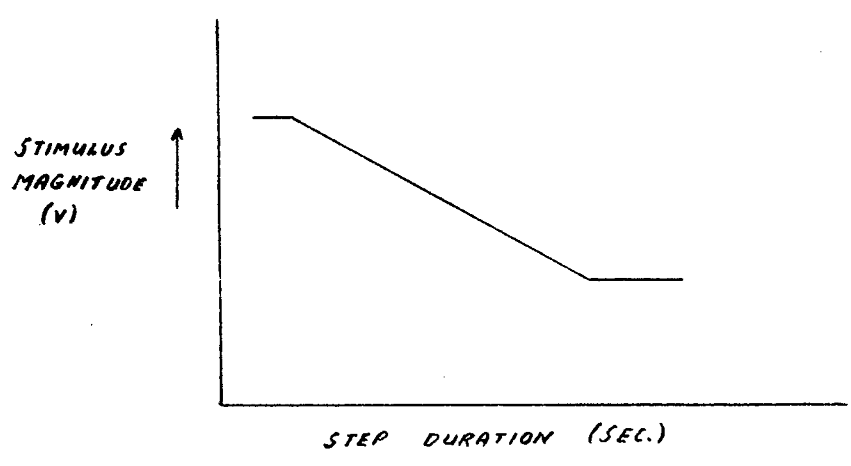


Fig. 10.6

10.6 STIMULUS SIGNAL.

The fixed parameters of the stimulus signal were taken from a consensus of other experimenters in this field and were set at a frequency of 50 Hz. with a pulse width of 0.3 msec. Square wave pulses were used.

CHAPTER 11.

ELECTRONIC CIRCUIT DESIGN.

11.1 TIME MEASUREMENT.

11.1.1 TIME SWITCH.

The pace period was measured by determining the length of time the toe of the unstimulated foot was on the ground. For this purpose a footswitch of the type described in 6.1.1.2, was attached to the toe, so as to form a closed circuit when toe-ground contact was made. This switch was more than adequate for the experimental model.

11.1.2 TIME BASE GENERATOR.

In order to store the pace period information and to activate the subsequent circuits according to its value, a voltage time base generator was used. This produced a voltage proportional in magnitude to the period of toe-ground contact.

Several circuits exist for accomplishing this time to voltage conversion. In this design, a Miller integrator was used due to the high degree of linearity which could be attained. The time base characteristic may be seen in Fig. 11.1.

The levelling off of the voltage after two seconds is a safety device which limits the duration and magnitude of the stimulus in case the patient should remain standing on the time switch for any length of time.

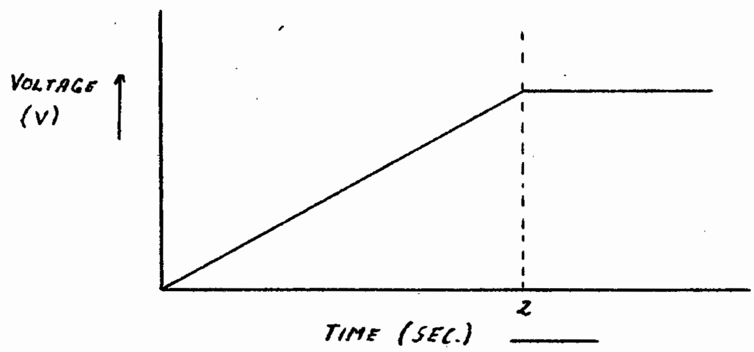


Fig. 11.1

Fig. 11.2 is the circuit of the Miller integrator.

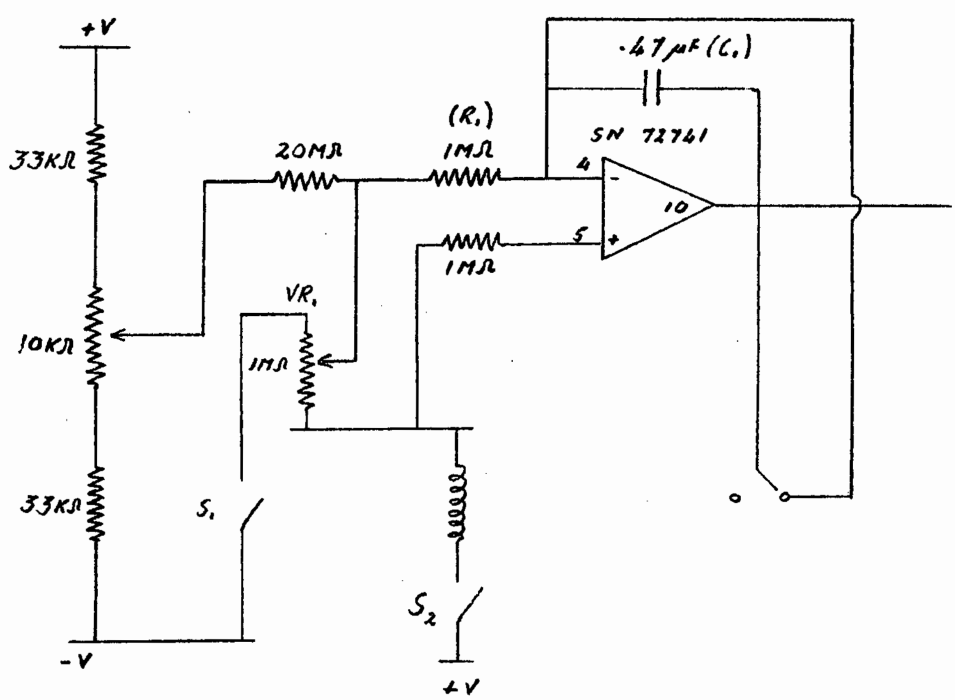


Fig. 11.2

When the toe-switch S is closed, a voltage appears at the input to the integrator and the capacitor C is charged up linearly with time. The charging speed (C.S.) is determined as follows:

$$\text{C.S.} = \frac{V}{R_1 C_1} \text{ volts/second.}$$

with $R = 1M\Omega$ and $C = .47\mu F$

C.S. = 2.13 V volts/second.

The changing time can thus be controlled by setting the variable resistor VR_1, which determines V. In order to achieve the curve of Fig. 11.1:

$V = .94$ volts.

The input to the integrator was also connected through a 20 MΩ resistor to a variable voltage. This was used to counteract a slight drift in the integrator output when switch S_1 was open.

The integrator was automatically discharged at the end of each cycle so that it would be ready for the following pace period measurement. This discharge was effected by a relay which short circuited the condenser C_1. The relay was activated by switch S_2 which was closed by the same toe-ground contact as S_1. The action was such that when switch S_2 was opened, capacitor C_1 was discharged. The relay was also responsible for the memory of the condenser voltage, but this will be discussed in a later section.

11.2 PULSE FORMATION.

An astable and monostable combination was used to set the frequency and width of the stimulus pulses.

11.2.1 FREQUENCY DETERMINATION.

The astable multivibrator circuit used was that shown in Fig. 11.3. This circuit produces square wave oscillations.

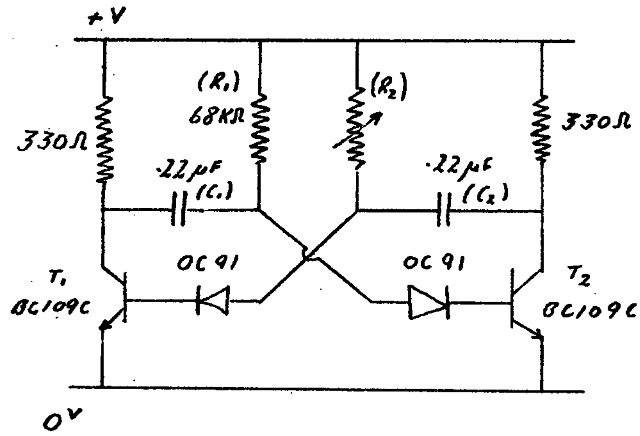


Fig. 11.3

The period of the stimulus waveform which is the inverse of the frequency may be determined as follows:

$$T = \frac{1}{f} = 0.69\ (R_1C_1 + R_2C_2)$$

$$= 0.69 \times .22\mu F\ (R_1 + R_2)$$

$$= 0.159 \times 10^{6}\ (R_1 + R_2) \text{ seconds.}$$

The frequency was set at 50 Hz. by fine setting of the resistor R_2. The two diodes in the base circuits of each transistor were to aid in quick transition of the transistor from one state to another. This helped to shorten the rise and fall times of the stimulus pulses.

11.2.2 PULSE WIDTH DETERMINATION.

The square wave pulse created by the astable multivibrator was fed into the monostable circuit of Fig. 11.4 for variable control of the pulse width.

The circuit of Fig. 11.4 is only half of a

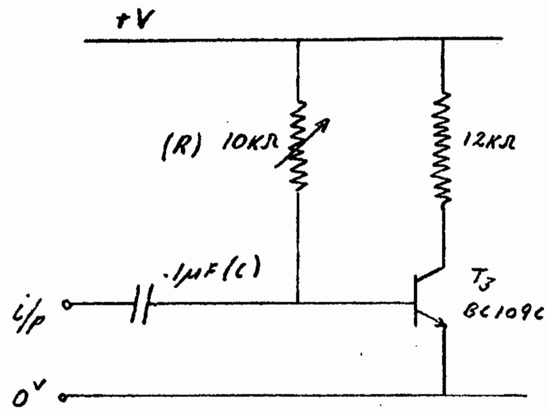


Fig. 11.4

normal monostable circuit, but the output transistor T_2 of the astable acts here as the other half of the monostable as well. As the collector of T_2 goes high, the base of T_3 goes high, driving it ON. T_3 then remains ON until the capacitor C has discharged through the resistor R.

The period of the output pulse is given approximately as:

$$T = 0.69 \text{ RC sec.}$$

In the final model this value was adjusted to:

$$T = 0.3 \text{ msec.}$$

11.3 THE STIMULUS DURATION CONTROL.

11.3.1 THE MEMORY.

Capacitor storage was used as the memory element for the duration control. A capacitor C_3 was charged up by the ramp generator of 11.1.2 when the switches S_1 and S_2 were closed. The capacitor C_3 was then disconnected from the ramp generator by the relay, when S_1 and S_2 were opened, and connected to the stimulus duration control circuit.

According to the system design in Chapter 10, the stimulus was to last 40% of the step period which is 80% of the toe-ground contact time. To do this, it was necessary to discharge the capacitor C_3 linearly so that it reached 0 volts after a time period equal to 80% of the ON time of switch S_1. This was achieved by discharging C_3 through a large resistor to the negative rail. The discharge characteristic down to 0 volts was thus reasonably linear. By varying the discharge resistance, the percentage of the step period for which stimulus was applied, could be controlled. A value of 2.8 MΩ was used to give the required stimulus duration.

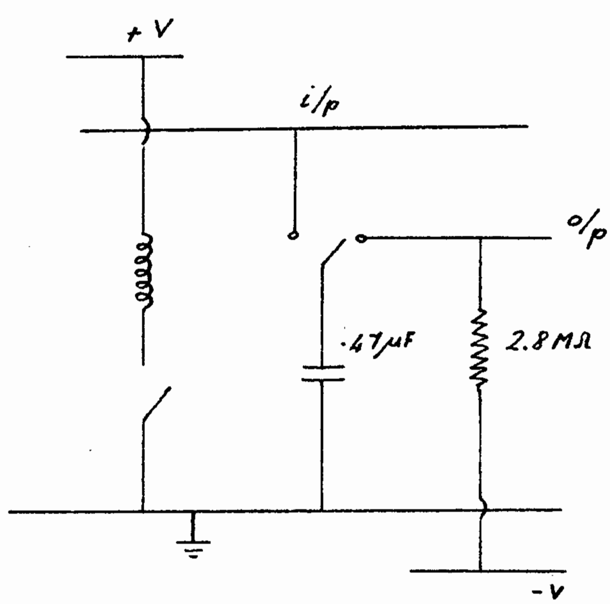


Fig. 11.5

The discharge characteristic may be seen in Fig. 11.6.

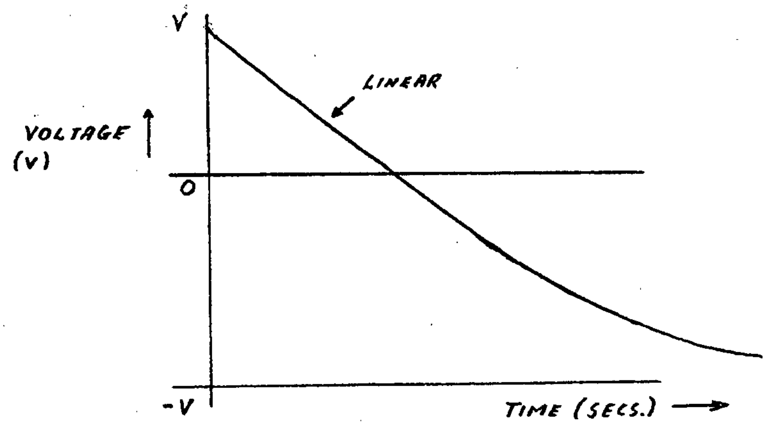


Fig. 11.6

11.3.2 VOLTAGE FOLLOWER.

A voltage follower was necessary at the output of the R.C. memory, since a minimum amount of current drain from the memory was essential to minimise distortion of the discharge waveform. This criterion was satisfied by the high input impedence of the follower.

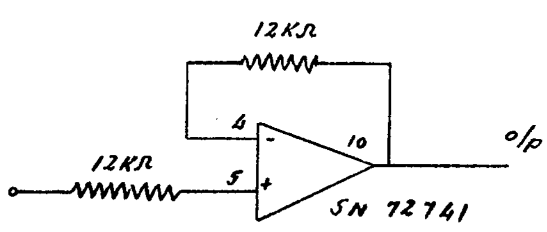


Fig. 11.7

11.3.3 SCHMITT TRIGGER.

The LM 710, an integrated circuit Schmitt trigger, was used to form a pulse of duration equal to the stimulus duration required. The Schmitt trigger was triggered negative when the input rose above 100 mV. and positive when it dropped below 0 V. Thus the pulse duration was equal to the time the voltage on capacitor C_3 remained above 0 V.

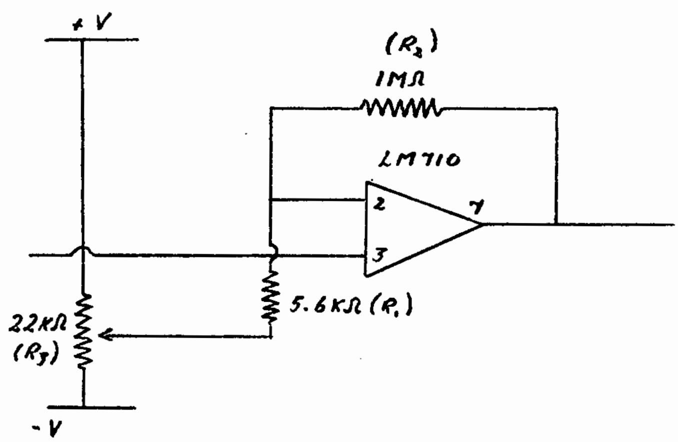


Fig. 11.8

By varying the ratio between R_1 and R_2 the differences between the triggering ON and triggering OFF voltages could be controlled. By varying the offset on resistor R_3 the actual voltages at which triggering occurred could be varied.

11.3.4 OUTPUT AMPLIFIER.

An output amplifier was used to invert the Schmitt trigger pulse and amplify it to a suitable level to control the FET switch of 11.3.5.

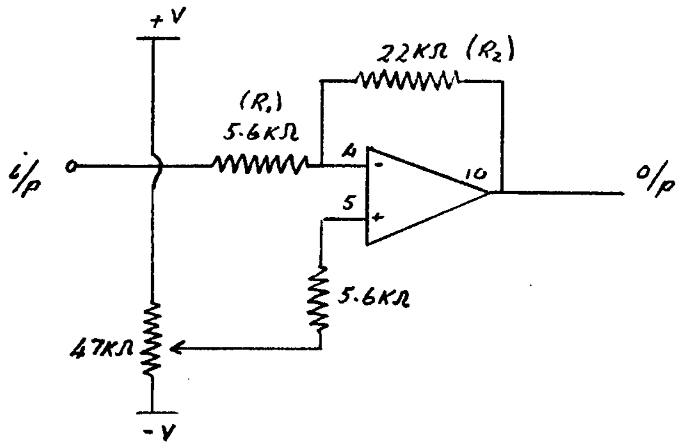


Fig. 11.9

$$\text{Amplification} = \frac{R_2}{R_1} = \frac{22K\Omega}{5.6K\Omega}$$
$$= 3.93$$

11.3.5 F.E.T. SWITCH.

An F.E.T. switch was used to control the passage of the stimulus pulses from the astable circuit and eventually to the stimulator itself. When the input to the gate of the F.E.T. went high the F.E.T. acted as a short circuit permitting the passage of pulses and when the gate voltage went low, the F.E.T. was an open circuit with respect to the 15KΩ resistor so that the square wave pulses no longer appeared at the output of the stimulator.

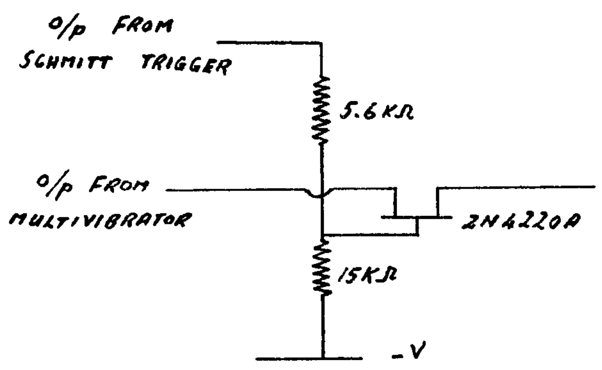


Fig. 11.10

11.4 THE STIMULUS AMPLITUDE CONTROL.

11.4.1 THE MEMORY.

A condenser memory was used here as in 11.3.1. Here, however, the relay simply separated the capacitor from the ramp generator and fed it into a voltage follower, with a very high input impedence so that the voltage would not leak away. The capacitor was not discharged as in 11.3.1.

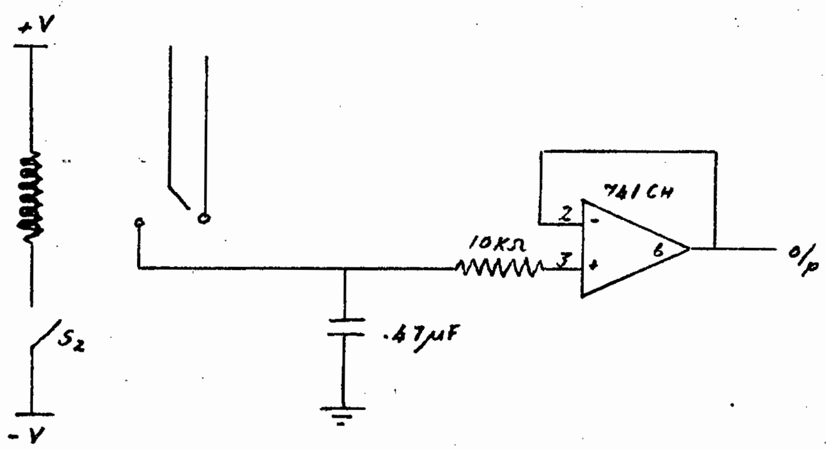


Fig. 11.11

11.4.2 VOLTAGE CONTROLLED AMPLIFIER.

The voltage controlled amplifier of Fig. 11.12 varies the amplification of the pulsed signal according to the value of the voltage at the base of transistor T_3.

As the voltage as the base of transistor T_3 increases, so the current through the transistor increases and the amplitude of the output pulse which appears at the collector increases. Thus as the input to T_3 increases, the amplification of the circuit increases. As this is the opposite to the desired condition layed out in Chapter 10, an additional transistor T_4 was added before T_3 to invert the output from the voltage controlling memory. Consequently, as the memory output increases the amplification of the circuit decreases.

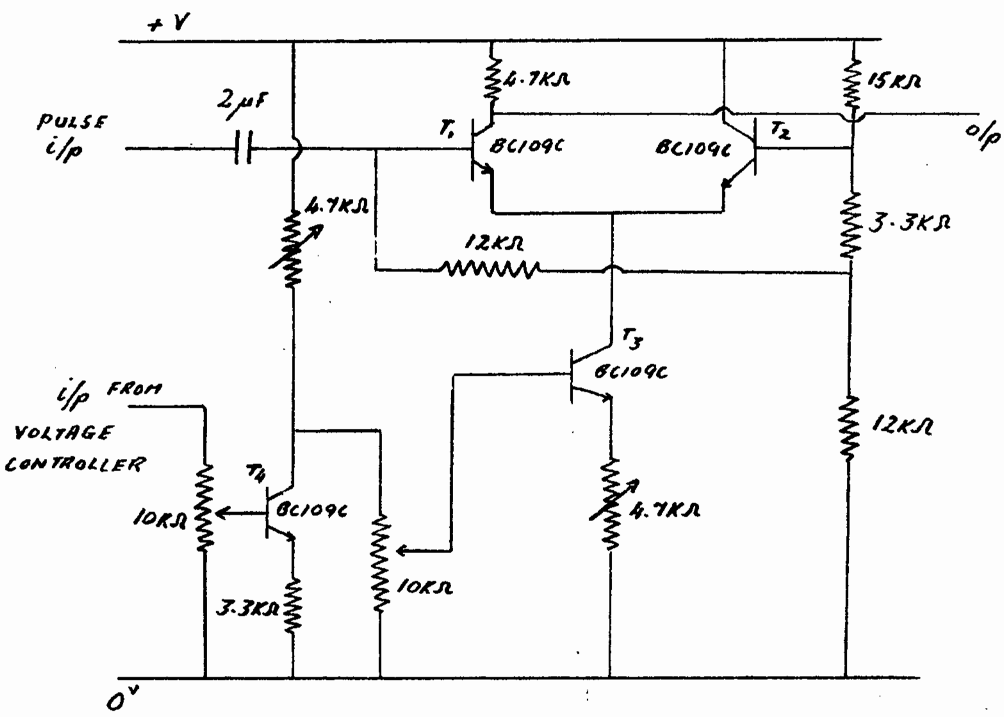


Fig. 11.12

11.5 OUTPUT AMPLIFIER.

An output amplifier was provided, firstly to invert the pulses from the voltage controlled amplifier, and also to set the amplitude of the pulses before they were fed onto the stimulator, since it was simpler to vary the amplitude here than on the stimulator itself.

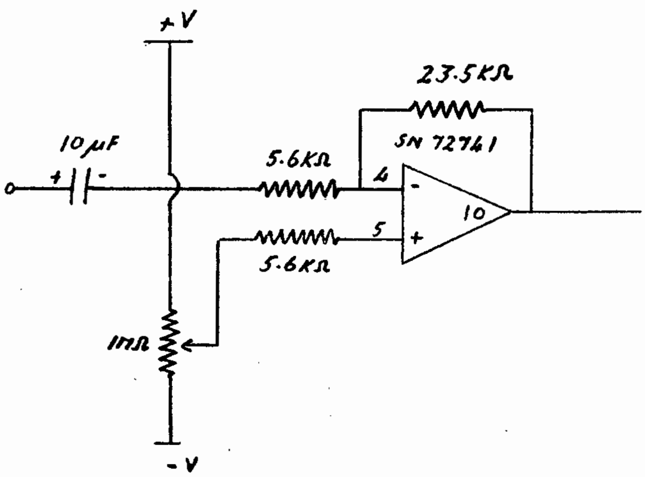


Fig. 11.13

11.6 THE STIMULATOR.

The output from the amplifier was fed into the base of a power transistor 2N 4240, which was connected across a 90 V. rail. The output was taken between the collector and the positive rail, thus supplying the stimulated area with a constant current stimulus.

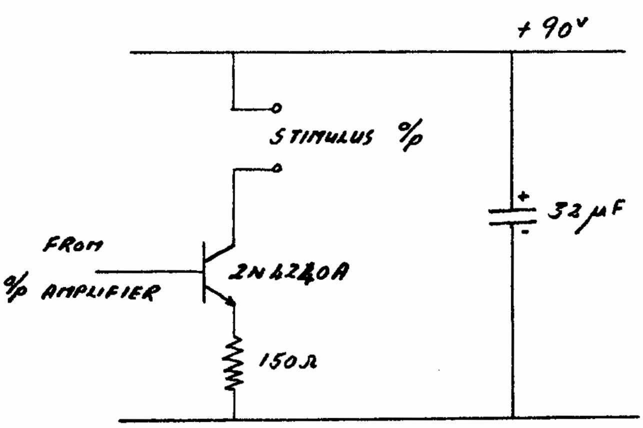


Fig. 11.14

11.7 ELECTRICAL IMPEDANCE BETWEEN ELECTRODES.

Although essentially it is not part of the stimulator design it is important to know the load impedance between the electrodes. The equivalent circuit in Fig. 11.15 was determined by Chandler and Sedgwick, (1971).

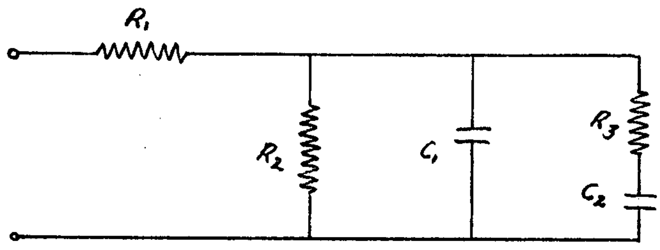


Fig. 11.15

R_1 is a series resistance due to the electrodes themselves. It is of the order of several hundred ohms.

R_2 is a parallel resistance present with electrodes at least a few square centimetres in area. It is in the order of 25KΩ.

C_1 is the capacitive effect across the electrodes. It is in the order of .1μF.

R_3 and C_3 are the resistance and capacitance effect presented by the skin itself, and are in the order of 2KΩ and .05μF respectively.

11.8 COMPLETE CIRCUIT DIAGRAM.

CHAPTER 12.

RESULTS.

12.1 STIMULATOR ELECTRODES.

The choice of stimulator electrodes was made after referring to an investigation by Milner et al (1970) into optimum electrode size and type. Firstly, for the earth electrode a fine stainless steel mesh, 7 cm. by 7 cm. was covered with Redux electrode paste and applied onto the tendon of the gastrocnemius muscle, about 5 cm. below the actual muscle bulk. This electrode was held in place with micropore paste. Whereas Milner et al used a small 3 cm. by 3 cm. mesh for the negative electrode and a button electrode covered with saline soaked gauze as the search electrode, the button electrode was used here as a permanent fixture for reasons which are mentioned below. The position of application of the button electrode was in the popliteal crease, just behind the knee, where it was shifted carefully till the correct position was found. In this position, stimulation of the nerves feeding the gastrocnemius muscle heads was achieved.

12.2 TESTING THE EQUIPMENT.

In order to evaluate the system in the most efficient manner, the following order of testing was used.

Before actually using the stimulator on a patient it was tested out on a normal person. This permitted greater freedom in the initial adjustments. A Grass Isolated Stimulator type 2533 was used in conjunction with a Grass Gated Pulse Generator type 2521 to stimulate the muscle initially, in order to determine the range of stimulus magnitudes required. This information was then used to set the final amplifier in the designed stimulator so that suitable stimulus levels were obtained.

After completing the initial testing on a normal person, the above process was repeated on a hemiplegic patient.

12.3 DIFFICULTIES ENCOUNTERED.

Several problems were encountered which made the gastrocnemius more difficult to stimulate than the tibialis anterior muscle, which was stimulated in the peroneal brace. Firstly, since the stimulation point was embedded deep in the back of the knee, it proved rather difficult to find and once found, could easily slip away, due to the lack of a hard surface upon which to catch the nerve.

Another difficulty encountered, once again due to the depth of the nerve amidst the tissue, was that a constant pressure was required upon the electrode. It was for this reason that it was decided to retain the button electrode as the permanent electrode for this stimulator. Care should be exercised, however, in seeing to it that no blood vessels are squeezed tight.

The button electrode alone was also found to supply insufficient pressure to the stimulation site and so a mould of the back of the knee was made. This was done by simply placing "Orthoplast" in a bowl of very hot water at which it becomes malleable and can be moulded to any shape. This mould was pressed against the button electrode and held in place with micropore tape. Under these conditions good results were obtained.

12.4 RESULTS.

In the initial testing of the apparatus on a healthy person, the system was found to blend in well with normal walking, and the timing of the stimulus was felt to coincide very closely with normal muscle activity.

The hemiplegic patient on whom the stimulator was tested, was intelligent and co-operated well. Certain difficulties, however, presented themselves. Firstly, as mentioned in 12.3, great difficulty was experienced in locating the stimulating point of the gastrocnemius muscle. If this system were to be used eventually as a regular functional aid, a simpler method of locating would have to be perfected.

Secondly, it would require a certain amount of training with the stimulator before the patient's abnormal walk could be made to co-ordinate with the stimulus frequence of the stimulator. A reasonable training period on a group of hemiplegics would thus be necessary in order to evaluate the overall possibilities of the instrument.

12.5 CONCLUSION.

Though not many hemiplegic patients suffer from loss of gastrocnemius activity alone, the use of this stimulator in conjunction with a peroneal brace in the more seriously affected hemiplegic cases, could be of great assistance. It is felt, however, that the real advantage of this system would be in evaluating the potential of functional electrical stimulation on a wider scale. A long-term study of this stimulator on a patient would be necessary before it could be definitely determined whether there was an improvement in the patient's gait pattern or not.

CHAPTER 13.

CONCLUSION.

The scope of functional electrical stimulation in the aid of the paralysed is extremely wide. Only the surface has been touched. Perhaps it might one day be possible for completely denervated legs to walk artificially. However, far more knowledge and research into the mechanism of gait control is necessary before this can become an actuality. It should be realised, though, that at the present moment, research is being done in nerve surgery, which might open up a completely different approach to rehabilitation work. The problems involved in this approach are, however, tremendous, and until some definite results are available, functional electrical stimulation seems to offer the most realistic and effective answer to a very pressing problem.

APPENDIX A.

A DIFFERENTIAL E.M.G. AMPLIFIER WITH ONE INPUT GROUNDED.

The differential amplifier circuit design appears below in Fig. A.1.

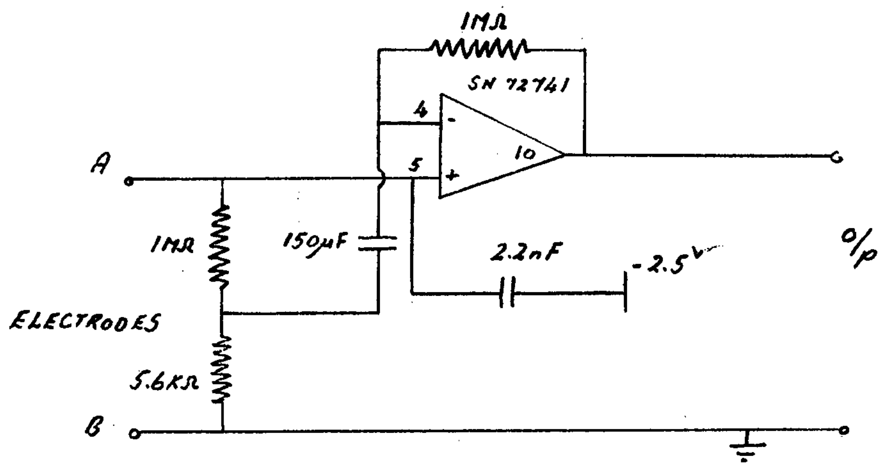


Fig. A.1

The Common Mode rejection ratio of the above circuit is theoretically infinite. However, since the two inputs are not completely isolated from earth, a certain amount of crosstalk occurs if extra e.m.g. amplifiers are coupled to the same earth. This is due to the fact that the amplifier registers the effect of the A electrode, not only with respect to its B electrode, but with respect to all other B electrodes as well. An experiment was performed to determine the extent of this interference and to decide whether the system ought to be used or not.

EXPERIMENT: To determine the crosstalk between two muscles using the above e.m.g. amplifier.

METHOD: 1. A square wave stimulus was applied across the biceps at a frequency of 200 Hz. with a pulse duration of 1 msec. The amplitude was then varied in steps.

2. The magnitude of the stimulus was measured by placing two monitoring electrodes between the stimulus electrodes.

3. The e.m.g. of the wrist flexors of the forearm were also monitored and amplified by 100.

RESULTS:

A. <u>Without stimuli applied to the biceps</u>. The wrist was flexed tightly and the e.m.g. magnitude was measured.

1. With no earth electrodes attached,

   e.m.g. magnitude = .3 - .5 volts.

2. With biceps earth electrode attached,

   e.m.g. magnitude = .3 - .5 volts.

B. <u>With stimulus applied across the biceps</u>. The wrist here was not flexed and thus only the interference was monitored by the forearm e.m.g. amplifier.

1. With no earth electrodes attached.

Size of stimulus (monitored at biceps)	Size of forearm amplifier output
10 mV.	0 V.
100 mV.	0 V.
1 V.	0 V.
10 V.	25 mV.

2. With biceps earth electrode attached.

Size of stimulus (monitored at biceps)	Size of forearm amplifier output
10 mV.	5 mV.
50 mV.	20 mV.
100 mV.	40 mV.
500 mV.	200 mV.
1 V.	.36 V.
10 V.	2.5 V.

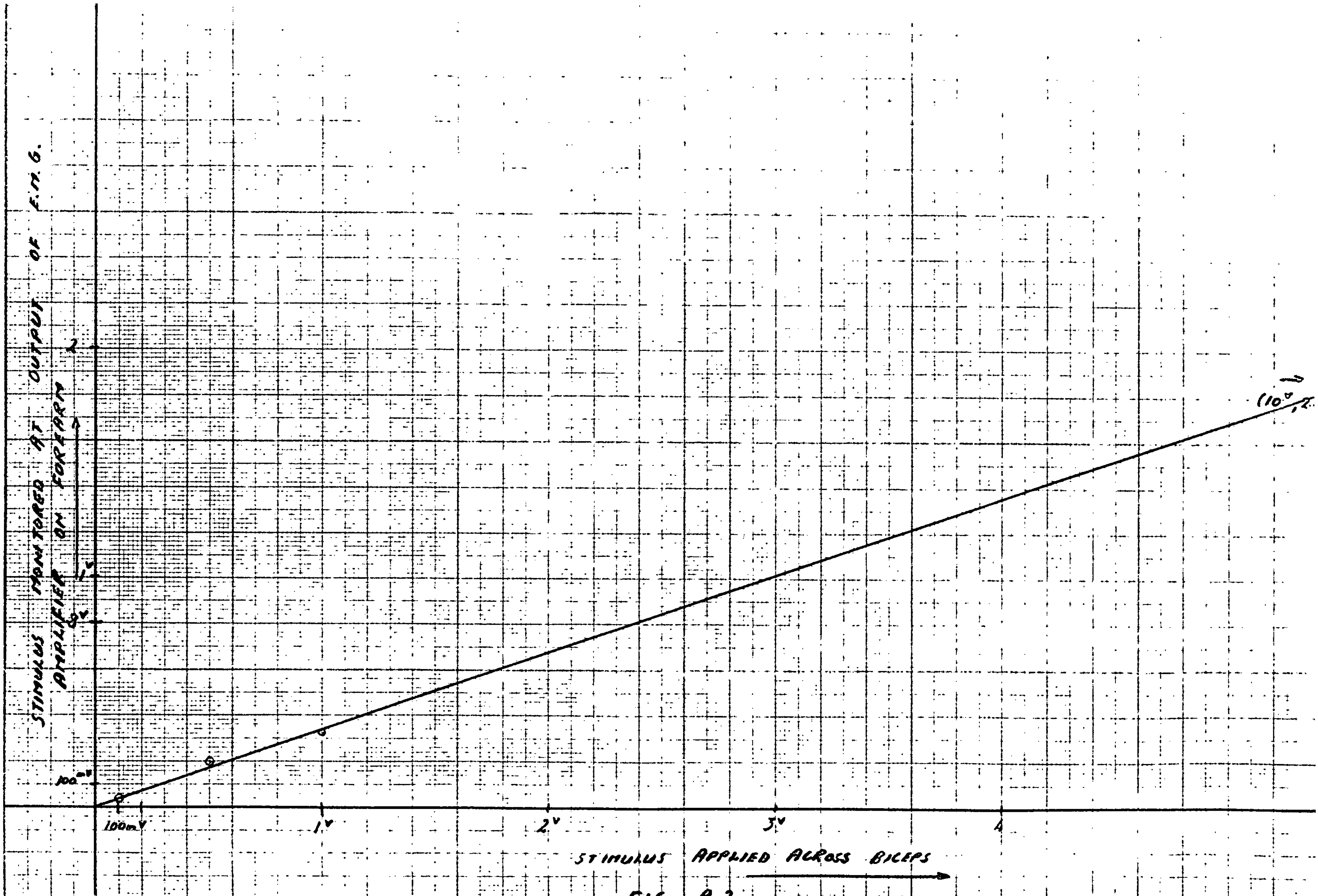
DISCUSSION:

The first fact noticeable from the above information is the additive effect of increasing the stimulus. The interference on the forearm amplifier is linearly proportional to the size of the stimulus.

For a 10 mV. biceps e.m.g. the crosstalk on the forearm amp would be 5 mV. As a percentage of the forearm e.m.g. output, this is:

$$\frac{5\text{mV.}}{.3\text{V.}} \times 100 = 1.66\% \quad \text{(maximum)}$$

However, when other channels were connected in a similar fashion to the same earth, the interference



STIMULUS MONITORED AT OUTPUT OF E.M.G. AMPLIFIER ON FOREARM
STIMULUS APPLIED ACROSS BICEPS
100mV
1V
2V
3V
4
100mV
.8V
1V
2
(10V, 2

FIG. A-2

was additive. Thus with a total of four channels, the maximum interference in any one channel was 5%.

CONCLUSION.

In a system which was completely free of other interference, 5% might be a tolerable limit. However, since interference was created in other parts of the system, it was decided against using these amplifiers.

APPENDIX B.

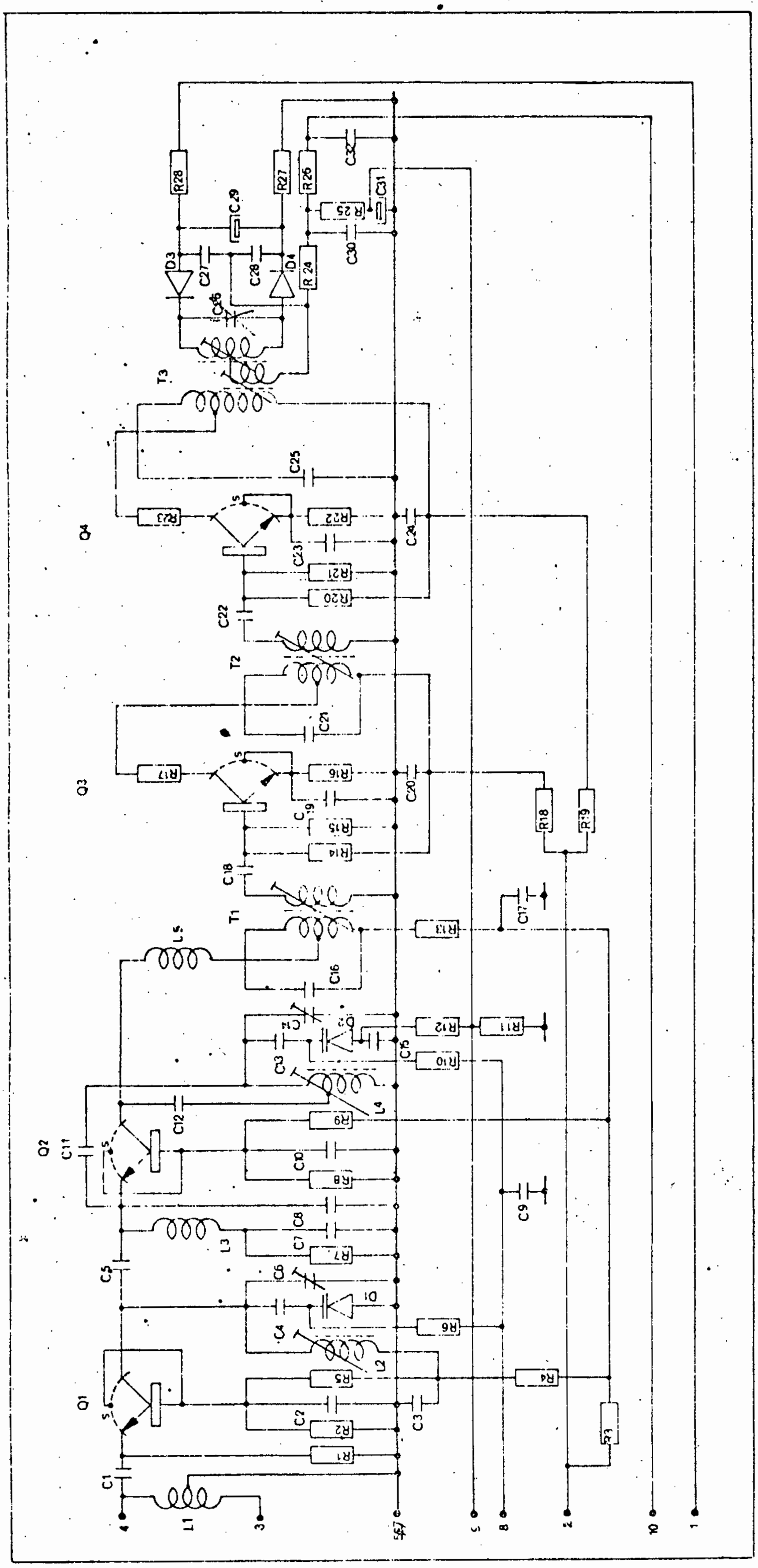
DETAILED CIRCUIT OF THE F.M. RECEIVER.

The F.M. receiver used was a commercially manufactured S.T.C. receiver available in modular form.

The circuit consisted basically of five sections:

1. Radio frequency (R.F.) amplifier.
2. Self mixing oscillator.
3. First intermediate frequency (I.F.) amplifier.
4. Second intermediate frequency amplifier.
5. Ratio detector.

A detailed analysis of the design was available with the receiver. This incorporated the circuit design as well, which is reproduced here.


Q1
Q2
Q3
Q4
T1
T2
T3
L1
L2
L3
L4
L5

Printed Circuit Board connection key.

With reference to Figures 10 and 11, the connector tracks of the module may be identified as follows.

1. Via a 200 microamp (F.S.D.) Tuning meter to earth (pins 5, 6 and 7).
2. Plus 5.6 v D.C., stabilised.
3. Antenna.
4. Antenna.

5, 6
& 7 Earth.

8. Tuning voltage from wiper of tuning potentiometer.
9. A.F.C. to earth via a switch.
10. Audio frequency output from module.

Component Complement

With reference to Figure 11.

Transistors

Q1, Q2: BF125 STC (SA)
Q3, Q4: BF121 STC (SA)

Diodes

D1, D2 : BB142 Variable capacitance diodes STC (SA)
D3, D4 : AA143 Germanium goldbonded diodes STC (SA)

Resistors

R1 - 680 ohm ¼ watt minimum 10% 470Ω
R2 - 3K3 ohm ¼ watt minimum 10% 1.5KΩ
R3 - 100 ohm ¼ watt minimum 10%
R4 - 100 ohm ¼ watt minimum 10%
R5 – 10K ohm ¼ watt minimum 10% 3.9KΩ
R6 – 56 Kohm ¼ watt minimum 5%
R7 – 680 ohm ¼ watt minimum 10% 470Ω
R8 - 3K3 ohm ¼ watt minimum 10% 1KΩ
R9 - 10K ohm ¼ watt minimum 10% 3.9KΩ
R10 - 56 Kohm ¼ watt minimum 5%
R11 10 Kohm ¼ watt minimum 10%
R12 1 Kohm ¼ watt minimum 10%
R13 - 10 ohm ¼ watt minimum 10%
R14 – 10 Kohm ¼ watt minimum 10% 27KΩ
R15 - 3K3 ohm ¼ watt minimum 10% 10KΩ
R16 – 470 ohm ¼ watt minimum 10%
R17 - 470 ohm ¼ watt minimum 10% 1KΩ
R18 – 330 ohm ¼ watt minimum 10%
R19 - 330 ohm ¼ watt minimum 10%
R20 – 10 Kohm ¼ watt minimum 10% 27KΩ
R21 – 3K3 ohm ¼ watt minimum 10% 10KΩ
R22 – 470 ohm ¼ watt minimum 10%
R23 – 470 ohm ¼ watt minimum 10% 1KΩ
R24 – 100 ohm ¼ watt minimum 10%
R25 – 22 Kohm ¼ watt minimum 10%
R26 – 4K7 ohm ¼ watt minimum 10%
R27, 28 – 18 Kohm ¼ watt minimum , 5%

Capacitors

C1 – 100pf. 12V ceramic disc capacitor
C2 – 0.001 uf 12V ceramic disc capacitor
C3 – 0.2 uf 12V ceramic disc capacitor .01μF
C4 – 0.005 uf mylar capacitor
C5 – 4 pf. 12V ceramic disc capacitor
C6 – 1.8 pf to 6 pf trimmer capacitor (7 or 10 mm base matrix).
C7 – 680 pf. 12V ceramic disc capacitor 470pF
C8 – 68 pf. 12V ceramic disc capacitor
C9 – 0.2 uf. 12V ceramic disc capacitor .01μF
C10 – 0.001 uf 12V ceramic disc capacitor
C11 – 4 pf. 12V ceramic disc capacitor
C12 – 68 pf. 12V ceramic disc capacitor
C13 – 0.005 uf mylar capacitor
C14 – 1.8 pf. to 6 pf. trimmer capacitor (7 or 10 mm base matrix).
C15 – 0.005 up mylar capacitor
C16 – 65 pf capacitor (Interwall to T1)
C17 – 0.2 uf 12V ceramic disc capacitor .01μF
C18 – 330 pf. 12V ceramic disc capacitor .001μF
C19 – 0.02 uf 12V ceramic disc capacitor
C20 – 0.2 uf 12V ceramic disc capacitor .01μF
C21 – 50 pf capacitor (Interwall to T2)
C22 – 330 pf. 12V ceramic disc capacitor .001μF
C23 – 0.02 uf 12V ceramic disc capacitor
C24 – 0.2 uf 12V ceramic disc capacitor .01μF
C25 – 7.5/8 pf. 12V ceramic disc capacitor
C26 – 80 pf capacitor (Internal T3)
C27,28 – 330 pf capacitors
C29,31 – 4.7 uf. 10V Tantulum electrolytic capacitors (S.T.C. (S.A.)).
C30 – 0.001 uf. 12V ceramic disc capacitor
C32 – 0.01 uf. 12V ceramic disc capacitor

Inductors

L1 6 Turns. 22SWG. Copper enamelled wire 5mm diameter, centre tapped.

L2 3½ Turns. 22 SWG. Copper enamelled wire on plastic former with dust Iron slug. 18 KΩ

L3 15 turns silk insulated Litz, wound on 56 Kohm resistor as former.

L4 1½ turns 22 SWG. Copper enamelled wire on plastic former with dust Iron slug, tapped 2½ turns from bottom.

L5 5 uH choke, preferably variable.

I.F. Transformers

T1. First I.F. Transformer: with reference to figure 19 for C and L nomenclature.

C_1	primary capacitance	= 65 pF
Qu	Unloaded Q	> 80
Wire	:	0.1° UEW

Primary turns : 14, tapped 8 turns from bottom

Secondary turns : 3

T2. Second I.F. Transformer: Specifications as above with the exception of C21 = 50 pf.

T3. Ratio detector filter: with reference to Figure 1(a) for C and L nomenclature.

C_1	C25 referring to Figure 11 = primary capacitance	= 7.5 pF
Qu	Unloaded Q	> 80
C_2	secondary capacitance	= 80 pF
Wire	:	0.08° UEW for primary and feedback coil
	:	0.1° UEW for Bifilarly wound secondaryc coil.

Secondary : Bifilarly wound : 5 + 5 turns.

Tertiary : Feedback winding : 3 turns.

The author examined a number of I.F. transformers and ratio detector filters (ready wound), currently available on the market, and found the selection to be wide. Inductances ranging from 0.08 uH to 1.5 mH, with a large variety of core materials available. The case sizes ranged from a small 5.5 mm square by 8.6 mm high, upwards. In view of the large number of interconnections available, particularly where the ratio detector filter is concerned, the printed circuit board would, in some cases have to be redesigned in the region of the ratio detector filter, to cater for some particular application.

Miscellaneous

Printed circuit Board : STC (SA) A.L. 100-001--00

Tuning meter : 200 microamp F.S.D.

Zener diode : D.C. supply voltage stabilising diode. STC (SA) ZG 5.6

Tuning potentiometer :

Connector Socket :

Unique features of the BF121/125 RF transistor family.

The actual collector, emitter and base junctions of the transistors conform to the normal, very small, interdigitated structure of most modern high frequency transistors i.e. see Figure 12 below:

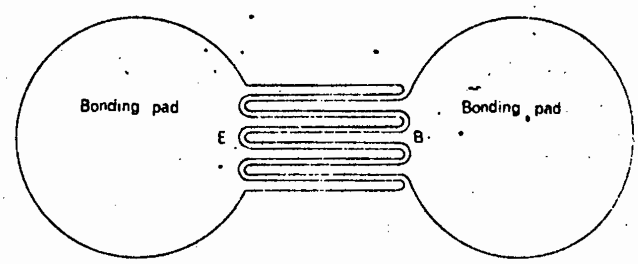


Figure 12

As a result of the minute sizes of the above "digits", the actual interjunction capacitances are extremely low. However, in order to apply the device, wires have to be bonded to the junctions of the transistor. Owing to the small size of the transistor, this is virtually impossible. For this reason, aluminium bonding pads, connected to the relevant junctions, are included in the structure of the transistor to facilitate the easy bonding of connection wires.

These bonding pads, however, give rise to significant collector base and collector emitter capacitances. In order to overcome this problem, diodes (P-type material) have been diffused into the N-type collector material beneath the bonding pads. Refer to Figure 13 below:

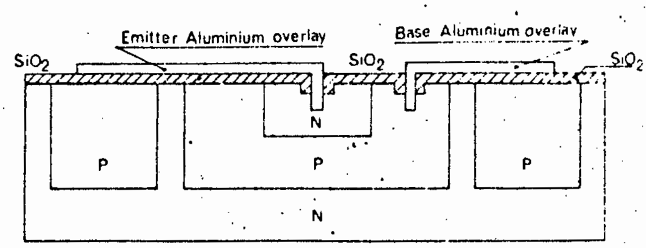


Figure 13.

These diodes effectively isolate the emitter and base aluminium overlay pads from the collector material, thus reducing interjunction capacitances to a minimum.

The two diodes are actually larger in plan area, than the emitter and base aluminium overlay pads thus minimising further, capacitive interactions due to fringe effects. The bodies of the two diodes are brought out from under the above mentioned overlay bonding pads and interconnected via a thin aluminium overlay structure which forms, as well, the bonding pad for the screen terminal of the transistor (see Figure 14) :

APPENDIX C.

THE DESIGN OF ACTIVE FILTERS USING TABLES.

An article by M. Bronzite (1970) which contains a set of tables for the simplified design of active filters, and details on how to use them, is reproduced here.

Simple Active Filters

Design procedure

by *M. Bronzite*, B.Sc.

In recent years there has been much work on low-frequency active filters using twin-tee, op-amps, n.i.c.s, and gyrators. For all of these, the calculation of the necessary frequency selective components can be tedious, and some knowledge of filter theory is desirable in order to match the chosen type of filter to the particular requirement. It is, perhaps, time to re-examine a simpler structure using unity-gain amplifiers[1,2], which lends itself to rapid design without the use of precision components, yet is stable and may be readily "bread-boarded".

This design of a low- or high-pass filter will rely on evaluating three dependent variables, any two of which may be used to determine the third: (1) the pass-band ripple (m dB), which constitutes the variation in output over the whole of the pass-band with a constant amplitude input; (2) the reject-band attenuation, one useful measure of this being the attenuation one octave away from the pass-band limit; and (3) the order of the filter (N) which is the number of filter elements required to achieve a given performance. Given, say, (1) and (2), this article will describe how the rest of the design may be accomplished.

The filter itself consists of simple units which are added together to provide the required complexity, and these units are shown in Fig. 1 along with the pertinent design equations. With types (a) and (d) the first set of components (R_1C_1) may be designed independently of the second set (R_2C_2), whereas in types (b) and (e) the series elements are equal in value, giving an advantage of one less active element being used at the cost of reduced component flexibility. Due to the amplifier isolation, each unit can be considered without regard to the requirements of other units and can even be separated from them by intervening linear circuitry without degrading the overall performance. In many cases, a value of C is chosen and the value of R is calculated on the grounds of restricted capacitor availability, and this tends to favour the use of units (a) and (e) for low- and high-pass filters respectively, since (b) requires two capacitor values and (d) requires two amplifiers. The unity gain amplifiers can consist of any available active element with a gain of $1 \pm 0{\cdot}05$ assuming the filter performance is not required to be too stringent. (Naturally, a very "tight" specification would demand both precision components and an accurate amplifier). Thus op-amps and emitter followers are of immediate application but some care must be taken with the design of source and cathode followers since their transmission characteristics can be significantly less than 0·95. The drive capability will depend on the source and load presented to the amplifier; i.e., using unit (d) from Fig. 1, if R_2 is much larger than R_1, then a Darlington pair would be used for the second amplifier, but if R_2 is very roughly equal to or smaller than R_1 then a simple emitter follower is suitable.

Now a filter pass-band limit may be defined as either the frequency at which the output has diminished by m dB (f_m) or the frequency where it has diminished by 3 dB (f_{3dB}) and obviously the attenuation in the first octave after this point will depend on which criterion is chosen. In the latter case, the filter performance is related to f_{3dB} and it is necessary to generate the equivalent value of f_m in order to apply the design equations given in Fig. 1. This is done by means of a coefficient β which is given in Table 3 for various values of ripple and order of filter, and the appropriate conversion equations are appended to the table. The calculation of β itself is derived from ref. 3.

The only matter outstanding to finish the design is the value of T_n and this is given in Tables 1 and 2, with an outline of its derivation given in the appendix. The tables contain nine groups of figures of which the first eight generate a Chebychev response ($m \neq 0$) and the last one generates a Butterworth response ($m = 0$ and $f_m = f_{3dB}$). The figures quoted in the attenuation column cater for the two different cases discussed above, and it would seem practical to use the first when m is large and the second when m is small. In any case, these attenuation figures were extrapolated from graphical sources[1,4,5] and can only be considered as approximate with a maximum error of $\pm 5\%$ on the quoted figure. While on the subject of attenuation it should be recalled as a rough rule of thumb that all the filters have a roll-off of $6N$ dB/octave after the first octave. Thus a five element 1-dB low-pass filter with a pass-band limit of 1 kHz will be 1 dB down at 1 kHz, 45 dB down at 2kHz (from Table 1), 75 dB down

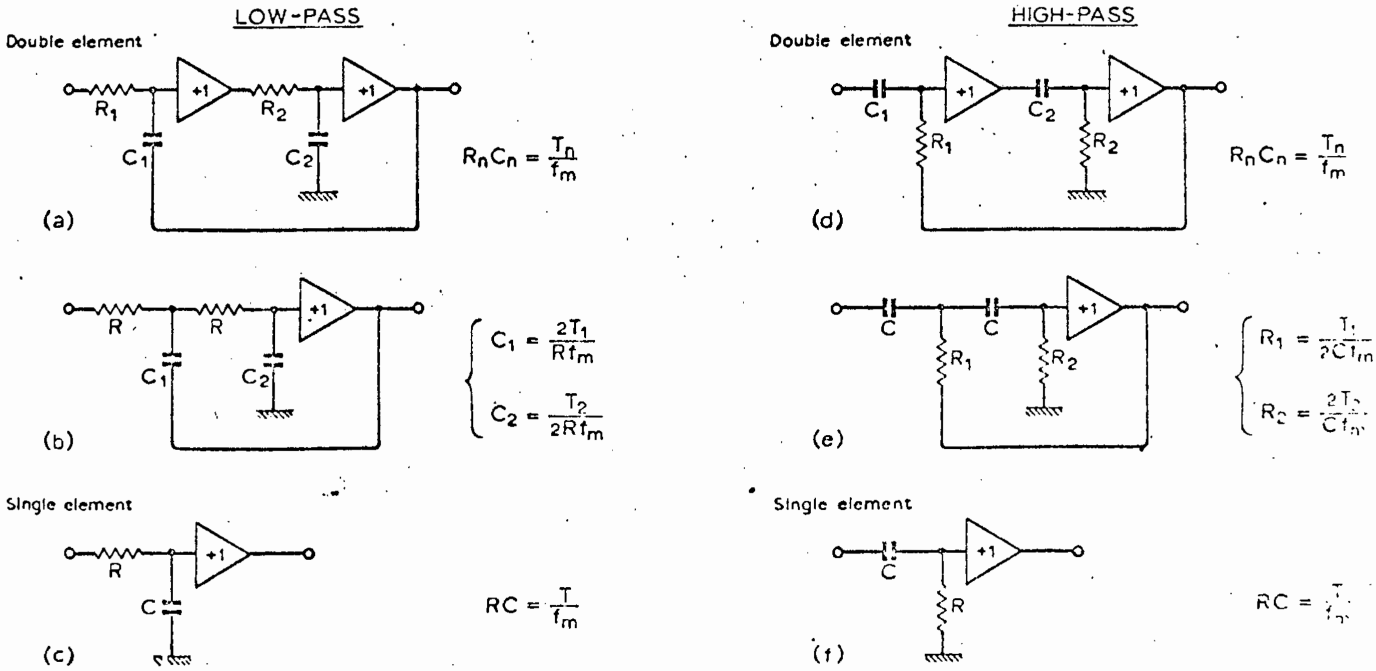


Fig. 1. Block configurations.

it 4 kHz (45 + 6 × 5), and so on. For more accurate figures, refs. 1 and 4 may be consulted, although the values given in the tables will be found adequate in the majority of case.

Having covered the process of design, two examples will be given to illustrate the approach. The first concerns a low-pass filter with a maximum permitted in-band variation of 2%, f_{3dB} = 4·5 kHz, and the first octave attenuation must be in excess of 50 dB. Now 2% is approximately 0·2 dB so m = 0·1. Examination of Table 1 gives a value of N = 6 for 52 dB of attenuation. Moving to Table 3, for the given values of m and N it is found that β = 1·093, and this in turn gives f_m = 4·5/1·093 = 4·12 kHz. Returning to Table 1, T_1 = 0·69383 for the first Double . . . and the rest of the design is straightforward, having agreed on which unit to use. The second example will be worked out in full and consists of a high-pass filter with a pass-band ripple of less than 10%, f_m = 100 Hz, and 50 Hz rejection must be better than 35 dB. Selecting m = 0·5 (6%) gives the required order as N = 5 with 42 dB attenuation. It was arbitrarily decided to use a 0·1-μF capacitor throughout, and the filter would consist of two (e) units with one (f) unit. Thus, with T_n selected from Table 2, for the first unit, D_1, $R_1 = 0{\cdot}0356/(2\times0{\cdot}1\times10^{-6}\times100)$ = 1·78 kΩ, $R_2 = 2\times0{\cdot}736/(0{\cdot}1\times10^{-6}\times100)$ = 147·2 kΩ; for D_2, $R_1 = 0{\cdot}0933/(2\times0{\cdot}1\times10^{-6}\times100)$ = 4·66 kΩ, $R_2 = 2\times0{\cdot}129/(0{\cdot}1\times10^{-6}\times100)$ = 25·8 kΩ; and for the (f) unit $R = 0{\cdot}0577/(0{\cdot}1\times10^{-6}\times100)$ = 5·77 kΩ. The final circuit is shown in Fig. 2 where the resistors are 5% and the capacitors are 10% tolerance. As this is a high-pass filter it is a good practice to decouple the h.t. lines, although it is hardly ever necessary for the low-pass circuits. The performance is shown in Fig. 3, and owing to the use of a relatively high distortion input signal there was some 2nd harmonic breakthrough below 30 dB which reduced the effective accuracy of measurement.

With the design established, some of the limitations of the filter will now be discussed and these should be borne in mind when considering a given filter for a given application. In the first place, no mention has been made of the pulse response of these filters and in general it can be said that the higher the ripple, and the higher the order, the more the overshoot on the output to a square-wave input. Where the matter is critical then Thomson filters[6,7] should be used, and using say, the values given in ref. 7, and applying the method given in the Appendix, values of T_n suitable for a maximally-flat delay filter may be readily found. On a more mundane subject care must be taken that the input amplitude does not approach that of the h.t. supplies. Apart from the problem that the emitter followers will have a large variation in output current (this can be minimized by using constant-current generators as emitter loads), amplification occurs in the heart of the filter, especially near the pass-band limit, which is not seen either at the input or output. Again, the higher the ripple, and the higher the order, the more the gain, and in practice, gains in the order of 6 dB or more may be

TABLE 1
Low-pass coefficients

Ripple order	Elements	Att. 1st octave		D1		D2		D3		Single
m dB	N	m dB	3 dB	T_1	T_2	T_1	T_2	T_1	T_2	T
3·000	2	17	17	0·24679	0·14498					
	3	28	28	0·53297	0·05664					0·53297
	4	39	39	0·93434	0·03002	0·38701	0·33397			
	5	51	51	1·45056	0·01866	0·55407	0·12126			0·89650
	6	62	62	2·08158	0·01274	0·76191	0·06371	0·55776	0·51140	
	7	75	75	2·82735	0·00927	1·00907	0·04002	0·69830	0·17759	1·25829
2·000	2	14	16	0·19800	0·15543					
	3	26	27	0·43142	0·06626					0·43142
	4	38	37	0·75870	0·03595	0·31426	0·36378			
	5	48	49	1·17961	0·02255	0·45057	0·14299			0·72904
	6	60	60	1·69411	0·01548	0·62009	0·07665	0·45393	0·55843	
	7	73	72	2·30217	0·01129	0·82164	0·04852	0·56859	0·20976	1·02456
1·000	2	11	15	0·14499	0·15847					
	3	22	26	0·32207	0·07911					0·32207
	4	34	36	0·57030	0·04502	0·23623	0·38378			
	5	45	47	0·86955	0·02881	0·33978	0·17365			0·54977
	6	57	58	1·27977	0·01998	0·46843	0·09696	0·34291	0·59233	
	7	70	69	1·74096	0·01466	0·62134	0·06239	0·42998	0·25563	0·77480
0·500	2	8	14	0·11164	0·14965					
	3	19	24	0·25406	0·08727					0·25406
	4	30	34	0·45381	0·05248	0·18798	0·37808			
	5	42	44	0·71075	0·03441	0·27148	0·19570			0·43927
	6	54	55	1·02482	0·02416	0·37511	0·11445	0·27460	0·58755	
	7	67	66	1·39602	0·01786	0·49823	0·07511	0·34479	0·28938	0·62129
0·100	2	3	13	0·06709	0·11393					
	3	12	22	0·16418	0·09131					0·16418
	4	23	31	0·30125	0·06322	0·12478	0·32588			
	5	35	40	0·47785	0·04436	0·18252	0·21824			0·29533
	6	47	52	0·69383	0·03233	0·25396	0·14323	0·18591	0·51735	
	7	61	62	0·94915	0·02443	0·33875	0·09928	0·23442	0·32722	0·42241
0·050	2	2	12	0·05509	0·09839					
	3	10	21	0·13996	0·08858					0·13996
	4	21	30	0·26049	0·06523	0·10790	0·29955			
	5	33	39	0·41602	0·04728	0·15890	0·21876			0·25711
	6	45	50	0·60633	0·03510	0·22193	0·15075	0·16246	0·48102	
	7	57	60	0·83134	0·02683	0·29670	0·10721	0·20533	0·33048	0·36998
0·010	2	0·5	12	0·03572	0·06802					
	3	5	20	0·10014	0·07721					0·10014
	4	15	28	0·19368	0·06519	0·08023	0·24303			
	5	27	37	0·31514	0·05112	0·12037	0·20768			0·19477
	6	39	47	0·46410	0·03978	0·16987	0·15882	0·12436	0·40265	
	7	51	58	0·64039	0·03133	0·22855	0·12006	0·15816	0·32024	0·28500
0·005	2	0·1	12	0·02982	0·05762					
	3	3	19	0·08757	0·07137					0·08757
	4	12	27	0·17258	0·06366	0·07148	0·22170			
	5	24	36	0·28339	0·05168	0·10825	0·19980			0·17515
	6	36	46	0·41950	0·04107	0·15355	0·15905	0·11241	0·37299	
	7	48	56	0·58069	0·03280	0·20725	0·12339	0·14342	0·31121	0·25843
0·000	2		12	0·11254	0·22508					
	3		18	0·15916	0·15916					0·15916
	4		24	0·08613	0·29408	0·20795	0·12181			
	5		30	0·09836	0·25752	0·25752	0·09836			0·15916
	6		36	0·08238	0·30746	0·11254	0·22508	0·30746	0·08239	
	7		42	0·08832	0·28679	0·12763	0·19846	0·35762	0·07083	0·15916

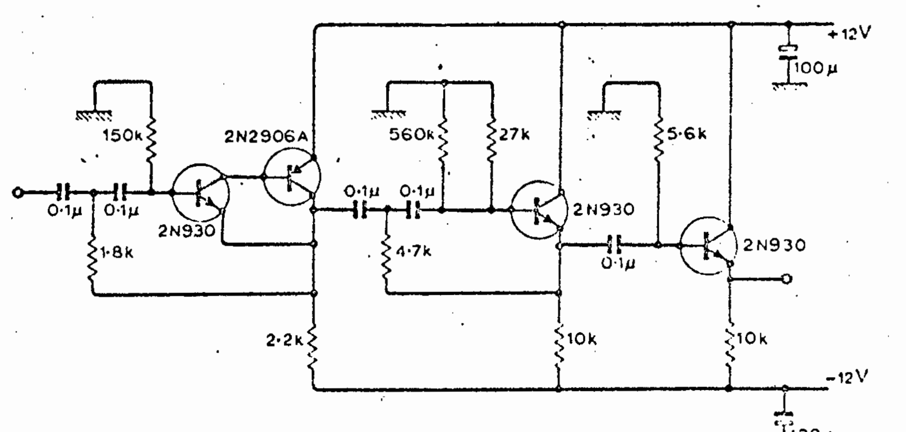


Fig. 2. Circuit of filter (see text).

encountered. However, an empirical approach will soon establish the extent of the problem and the permitted input levels for a given supply may be easily found. The choice of active element will depend to a certain extent on the frequency of operation envisaged. At the v.l.f. end, in order to keep the size of capacitors to reasonable proportions (and with exact requirements it is far easier to obtain low value precision capacitors), Darlington pairs of f.e.ts should be used which permit resistors in excess of 10 MΩ. At the h.f. end, high f_T transistors permit reliable operation up to, say, 10 MHz, in direct contradistinction to op-amp filters where 100 kHz represents a sensible limit. With this range, and using high density packaging for the active elements, video band-pass amplifiers without transformers or chokes become a distinct possibility. Again, d.c. offsets may dictate the selection of components; e.g., in a digital filter where

TABLE 2
High-pass coefficients

pple der	Elements	Att. 1st octave		D₁		D₂		D₃		Single
dB	N	m dB	3 dB	T_1	T_2	T_1	T_2	T_1	T_2	T
000	2	17	17	0·10264	0 17472					
	3	28	28	0·04753	0 44725					0·04753
	4	39	39	0·02711	0·84379	0 06545	0·07585			
	5	51	51	0 01746	1 35776	0·04572	0·20889			0·02825
	6	62	62	0·01217	1·98755	0·03325	0·39758	0·04541	0·04953	
	7	75	75	0·00896	2·73259	0·02510	0·63295	0·03627	0·14263	0·02013
000	2	14	16	0 12793	0·16297					
	3	26	27	0·05871	0·38228					0·05871
	4	38	37	0 03339	0·70458	0·08060	0·06963			
	5	48	49	0·02147	1·12319	0 05622	0·17714			0·03474
	6	60	60	0·01495	1·63643	0·04085	0·33047	0·05580	0·04536	
	7	73	72	0·01100	2·24373	0·03083	0·52206	0·04455	0·12076	0·02472
000	2	11	15	0·17471	0·15985					
	3	22	26	0·07865	0·32020					0·07865
	4	34	36	0·04442	0·56261	0·10723	0·06600			
	5	45	47	0·02848	0·87916	0·07455	0·14587			0·04607
	6	57	58	0·01979	1·26791	0·05407	0·26125	0·07387	0·04276	
	7	70	69	0·01455	1·72822	0·04077	0·40602	0·05891	0·09909	0·03269
500	2	8	14	0·22690	0·16927					
	3	19	24	0·09970	0·29025					0·09970
	4	30	34	0·05582	0·48264	0·13475	0·06700			
	5	42	44	0·03564	0·73618	0·09330	0·12943			0·05766
	6	54	55	0·02472	1·04842	0·06753	0·22132	0·09224	0·04311	
	7	67	66	0·01814	1·41851	0·05084	0·33725	0·07347	0·08753	0·04077
100	2	3	13	0·37757	0 22233					
	3	12	22	0·15429	0·27742					0·15429
	4	23	31	0·08408	0·40067	0·20300	0·07773			
	5	35	40	0·05301	0 57100	0·13878	0·11607			0·08577
	6	47	52	0·03651	0·78360	0·09974	0·17685	0·13625	0·04896	
	7	61	62	0·02669	1·03689	0·07478	0·25515	0·10806	0·07741	0·05997
)50	2	2	12	0·45981	0·25745					
	3	10	21	0·18098	0·28595					0·18098
	4	21	30	0·09724	0·38834	0·23476	0·08456			
	5	33	39	0·06089	0·53570	0·15941	0·11579			0·09852
	6	45	50	0·04178	0·72162	0·11414	0·16803	0·15591	0·05266	
	7	57	60	0·03047	0·94402	0·08537	0·23627	0·12337	0·07665	0·06846
)10	2	0·5	12	0·70912	0·37242					
	3	5	20	0·25296	0·32806					0·25296
	4	15	28	0·13078	0·38858	0·31574	0·10423			
	5	27	37	0·08038	0·49548	0·21043	0·12197			0·13005
	6	39	47	0·05458	0·63671	0·14911	0·15949	0·20369	0·06291	
	7	51	58	0·03955	0·80839	0·11083	0·21098	0·16015	0·07910	0·08888
005	2	0·1	12	0·84936	0·43959					
	3	3	19	0·28925	0·35493					0·28925
	4	12	27	0·14678	0·39787	0·35436	0·11426			
	5	24	36	0·08938	0·49034	0·23401	0·12678			0·14462
	6	36	46	0·06038	0·61675	0·16497	0·15926	0·22535	0·06791	
	7	48	56	0·04362	0·77218	0·12222	0·20528	0·17662	0·08139	0·09802
00	2		12	0·22508	0·12254					
	3		18	0·15916	0·15916					0·15916
	4		24	0·29408	0·08613	0·12181	0·20795			
	5		30	0·25752	0·09836	0·09836	0·25752			0·15916
	6		36	0·30746	0·08238	0·22508	0·11254	0·08239	0·30746	
	7		42	0·28679	0·08832	0·19846	0·12763	0 07083	0·35762	0·15916

TABLE 3
Value of coefficient β for f 3 dB filters

	m = 0·500	0·100	0 050	0 010	0·005
N = 2	1·390	1·943	2·268	2·667	3·903
3	1·167	1·359	1 512	1·728	2·075
4	1·093	1·213	1 278	1·390	1·566
5	1·059	1·135	1·175	1·245	1·351
6	1·041	1·093	1·121	1·168	1·240
7	1·030	1·068	1·088	1·122	1·175

Low-pass: $f_m = f_{3dB}/\beta$

High-pass: $f_m = f_{3dB}\ \beta$

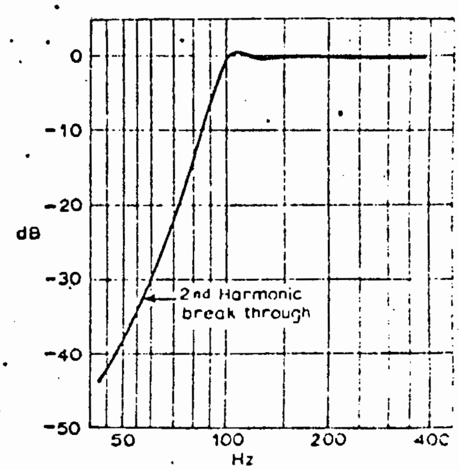


Fig. 3. Performance of filter

...ber of identical low-pass units are ...and any offsets would constitute a ...s noise problem. In this case, a first ...palliative would be to use p-n-p ...ating with n-p-n transistors for the ...d second amplifiers ("throwing in" ...ra emitter follower if N is odd), but ...is not good enough then it will be ...ary to revert to feedback amplifiers ...vide the unity gain.

...dix

...llowing analysis will indicate the way ...ch T_n has been calculated for Tables ...2, and will show how the method may ...d for creating other types of filters ...s Thomson). Considering unit (a) in

$$\left.\begin{aligned} 1/R_{in} &= 0 \\ R_0 &= 0 \\ \text{Gain} &= 1 \end{aligned}\right\} \text{for the amplifiers.}$$

= input voltage
= output voltage of second amplifier
= output voltage of first amplifier
= transmission function of unit

and let $\quad p_n = \omega C_n R_n$

then $\quad v_1 = \dfrac{v_{in}-v_0}{1+jp_1}+v_0 = \dfrac{(v_{in}+jp_1v_0)}{1+jp_1}$

and $\quad v_0 = \dfrac{v_{in}+jp_1v_0}{(1+jp_1)(1+jp_2)}$

i.e. $\quad v_0 = \dfrac{v_{in}}{(1+jp_1)(1+jp_2)-jp_1}$

or $\quad G = \dfrac{1}{-p_1p_2+jp_2+1}$

Putting $s = j\omega$ and $t_n = R_nC_n$

then $\quad G = \dfrac{1}{s^2t_1t_2+st_2+1}$

or $\quad G = \dfrac{1/(t_1t_2)}{s^2+s/t_1+1/(t_1t_2)}$

and similar expressions can be developed for the other double units. Now, any filter with zeroes at infinity can be expressed as

$$G = \left[(s^2+as+b)(s^2+cs+d)\ldots\right]^{-1}\times\alpha$$

where $\alpha = 1$ for low-pass filters and $\alpha = s^N$ for high-pass filters, and the values of a, b, c, d, ... can be found from the mathematical formulation of the filter under consideration. (Thus for a Butterworth two-element network, a = 1·414 and b = 1·000, while for a Thomson four-element network a = 5·792, and b = 9·140, and so on.)

Then, taking the first quadratic expression and equating coefficients,

$$a = 1/t_1$$
$$b = 1/(t_1t_2)$$

i.e.,
$$t_1 = 1/a$$
$$t_2 = a/b$$

But the above expressions are related to the angular frequency ω = 1, and must be converted to $f = f_m$, giving

$$t_1 = 1/(2\pi f_m a)$$
$$t_2 = a/(2\pi f_m b)$$

i.e., $R_1C_1 = T_1/f_m$ where $T_1 = 1/(2\pi a)$
$R_2C_2 = T_2/f_m$ where $T_2 = a/(2\pi b)$

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APPENDIX D.

PASSIVE FILTERS.

The following are calculated values for the passive filters used. The bandpass filters consisted simply of a high and a low pass filter cascaded.

The required bandpass sections were:

1. 8 KHz. - 12 KHz.
2. 17 KHz. - 21 KHz.
3. 29 KHz. - 33 KHz.

Also required were:

4. A high pass section for signals above 45 KHz.
5. A low pass section for 0 to 2 KHz.

A. Low pass filter sections.

The filters consisted of four sections.

a) Two end sections for matching.

b) The M-derived section for sharp cutoff.

c) The prototype section for maintaining the attenuation level after the attenuation due to b) decreases.

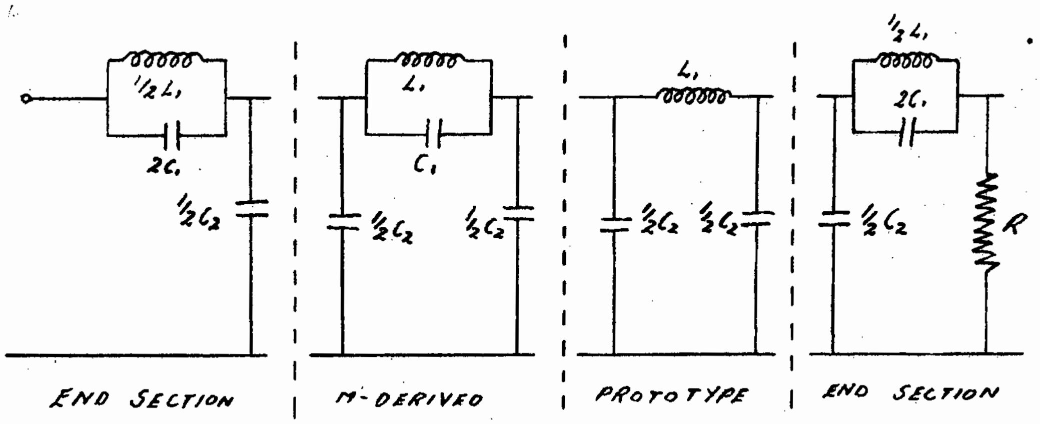


Fig. D.1

The first value to determine is m which is given as follows:

$$m = \sqrt{1 - (f_2/f_\infty)^2}$$

where f_2 is the frequency at which cutoff begins and f_∞ is the frequency at which maximum attenuation is desired.

For the prototype $m = 1$ } These are optimum values.
For the end section $m = 0.6$

With a desired f_2/f_∞ of .95, i.e. the maximum attenuation occurs within 5% of the cutoff frequency.

$$m = \sqrt{1 - (.95)^2}$$

$$= 0.3$$

This was the value of m for the M-derived circuit.

The formulae for determining the component values are as follows:

$$L_k = \frac{R}{\Pi f_2}$$

$$C_k = \frac{1}{\Pi f_2 R}$$

where R is the load resistance, set equal to 1KΩ.

Then:

$$L_1 = mL_k$$

$$C_1 = \frac{1 - m^2}{4m} C_k$$

$$C_2 = mC_k$$

1. <u>Low pass filter for 2 KHz</u>. ($m = 0.6$ for the M-derived section only in this case.)

a) End Sections.

$\frac{1}{2}L_1$ = 47.8 mH.

$2C_1$ = 84.88 nF.

$\frac{1}{2}C_2$ = 47.8 nF.

b) M-derived section.

L_1 = 95.5 mH.

C_1 = 42.44 nF.

$\frac{1}{2}C_2$ = 47.8 nF.

c) Prototype section.

L_1 = 159.2 mH.

$\frac{1}{2}C_2$ = 79.6 nF

2. Low pass filter for 12.5 KHz.

a) End sections.

$\frac{1}{2}L_1$ = 7.64 mH.

$2C_1$ = 13.56 nF.

$\frac{1}{2}C_2$ = 7.64 nF.

b) M-derived section.

L_1 = 7.64 mH.

C_1 = 19.28 nF.

$\frac{1}{2}C_2$ = 3.82 nF.

c) Prototype section.

L_1 = 25.45 mH.

$\frac{1}{2}C_2$ = 12.73 nF.

3. Low pass filter for 20.5 KHz.

a) End sections.

$\frac{1}{2}L_1$ = 4.66 mH.

$2C_1$ = 8.28 nF.

$\frac{1}{2}C_2$ = 4.66 nF.

b) M-derived section.

L_1 = 4.66 mH.

C_1 = 11.76 nF.

$\frac{1}{2}C_2$ = 2.33 nF.

c) Prototype section.

L_1 = 15.53 mH.

$\frac{1}{2}C_2$ = 7.77 nF.

4. Low pass filter for 34 KHz.

a) End sections.

$\frac{1}{2}L_1$ = 2.81 mH.

$2C_1$ = 5.0 nF.

$\frac{1}{2}C_2$ = 2.81 nF.

b) M-derived section.

L_1 = 2.81 mH.

C_1 = 7.11 nF.

$\frac{1}{2}C_2$ = 1.41 nF.

c) Prototype section.

L_1 = 9.37 mH.

$\frac{1}{2}C_2$ = 4.69 nF.

B. High pass filter sections.

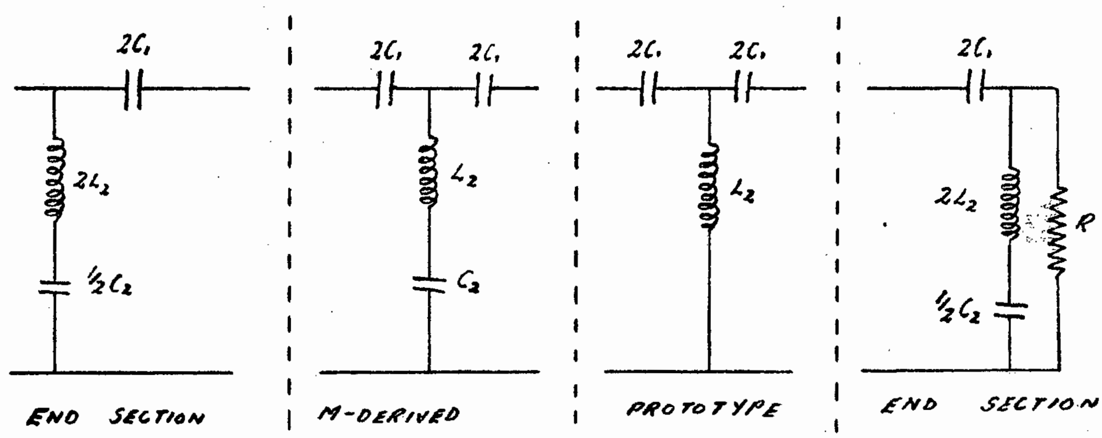


Fig. D.2

The filter has the same four sub-sections as for the low pass filter.

"m" for the high pass section is determined as follows:

$m = \sqrt{1 - (f_\infty/f_1)^2}$ where f_1 is the frequency at which cutoff begins and f_∞ is the frequency at which maximum attenuation occurs.

Again, for the prototype $m = 1$

for the end section $m = 0.6$

for the M-derived section $m = 0.3$

The formulae for determining the component values are as follows:

$$L_k = \frac{R}{4\Pi f_1}$$

$$C_k = \frac{1}{4\Pi f_1 R}$$

where R is the load resistance set equal to 1K .

Then:

$$C_1 = \frac{C_k}{m}$$

$$C_2 = \frac{4m}{1-m^2} C_k$$

$$L_2 = \frac{L_k}{m}$$

1. <u>High pass filter for 8 KHz</u>.

a) <u>End sections</u>.

$2C_1$ = 33.16 nF.

$2L_2$ = 33.16 mH.

$\frac{1}{2}C_2$ = 18.7 nF.

b) <u>M-derived section</u>.

$2C_1 = 66.3$ nF.

$C_2 = 13.12$ nF.

$L_2 = 33.15$ mH.

c) Prototype section.

$2C_1 = 19.9$ nF.

$L_2 = 9.95$ mH.

2. High pass filter for 16 KHz.

a) End sections.

$2C_1 = 16.58$ nF.

$\frac{1}{2}C_2 = 9.33$ nF.

$2L_2 = 16.58$ mH.

b) M-derived section.

$2C_1 = 33.16$ nF.

$C_2 = 6.57$ nF.

$L_2 = 16.58$ mH.

c) Prototype section.

$2C_1 = 9.95$ nF.

$L_2 = 4.975$ mH.

3. High pass filter for 29.5 KHz.

a) End sections.

$2C_1 = 8.99$ nF.

$\frac{1}{2}C_2 = 5.06$ nF.

$2L_2 = 8.99$ mH.

b) M-derived section.

$2C_1 = 17.86$ nF.

$C_2 = 5.06$ nF.

$2L_2 = 8.99$ mH.

c) Prototype section.

$2C_1 = 5.39$ nF.

$L_2 = 2.695$ mH.

4. <u>High pass filter for 45 KHz</u>.

a) <u>End sections</u>.

$2C_1 = 5.9$ nF.

$\frac{1}{2}C_2 = 3.32$ nF.

$2L_2 = 5.9$ mH.

b) <u>M-derived section</u>.

$2C_1 = 11.78$ nF.

$C_2 = 2.33$ nF.

$L_2 = 5.89$ mH.

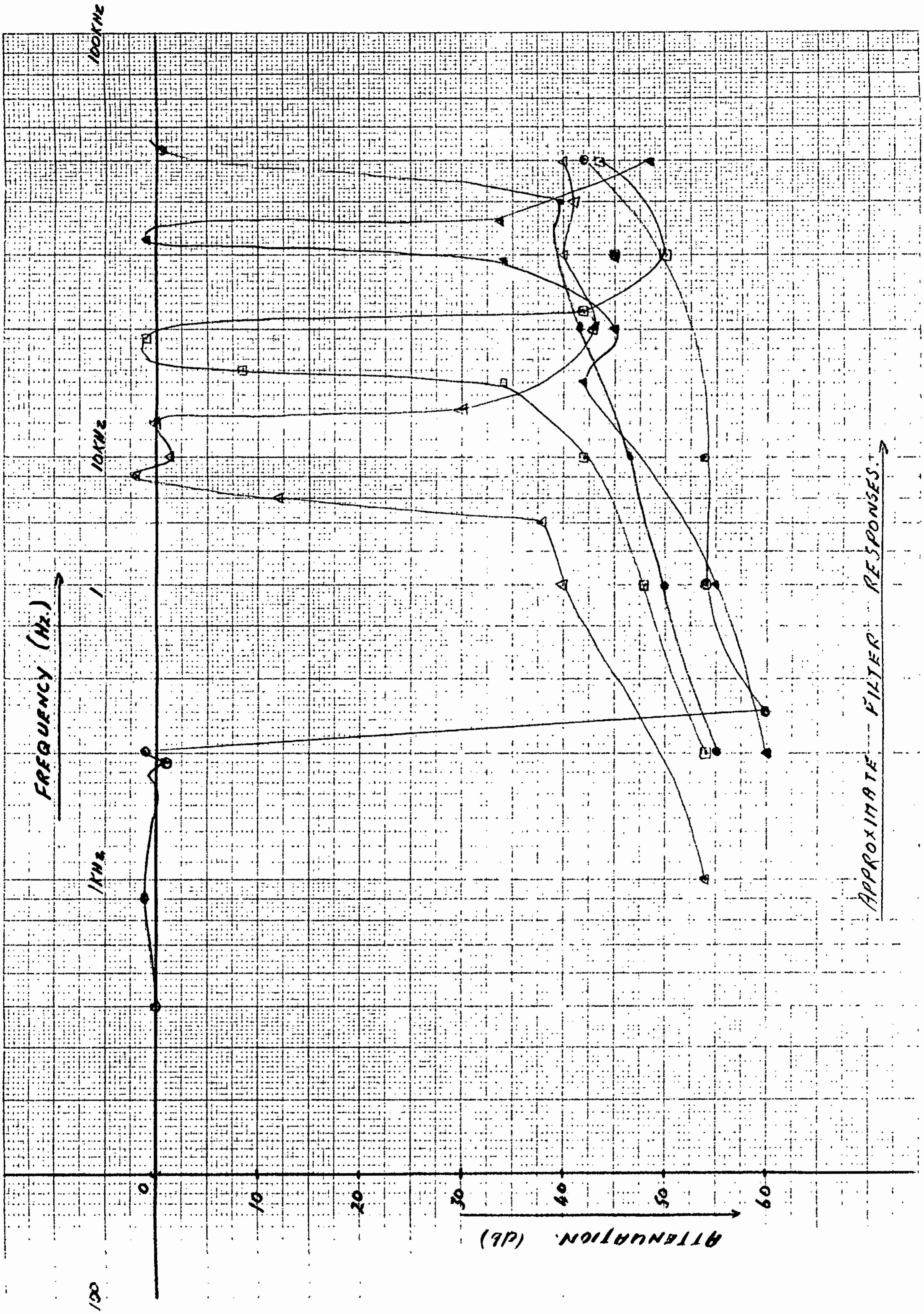
c) <u>Prototype section</u>.

$2C_1 = 3.536$ nF.

$L_2 = 1.768$ mH.

A graph of the filter characteristics may be seen in Fig. D.3.

HELIOS Log 3 Cycles x mm, 1/2 and 1 cm



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