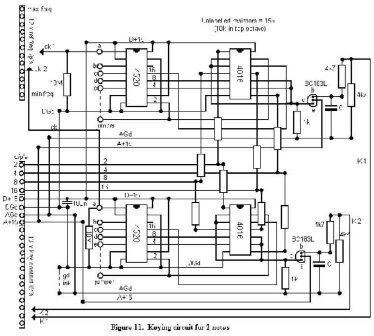
A COMPENDIUM OF ELECTRONIC ORGAN TECHNOLOGY

Volume 1: Analogue Organs

by C E Pykett



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by C E Pykett PhD FInstP

Considerable care has been exercised in the preparation of this volume. However the author cannot accept responsibility for any consequences howsoever they may arise.

Readers are cautioned against attempting to construct or use electronic equipment which involves mains voltages unless they are fully conversant with the health and safety aspects involved.

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ACKNOWLEDGMENT
Thanks are due to Dr A D Ryder for permission to reproduce the circuit of his CFP oscillator which first appeared in the Electronic Organ Magazine

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To Sandra

FOREWORD

This volume was written originally for members of the Electronic Organ Constructors' Society (EOCS) in 2001. It is a compendium of circuits and techniques which can be used in analogue electronic organs, and a second volume covers digital techniques. Comprehensive references to issues of the Electronic Organ Magazine (EOM – the house journal of the Society) and other publications are provided, and enough information is given for complete instruments to be built if desired using both divider and free phase methods.

Component availability has been a major preoccupation. For the divider and free phase tone generating systems only industry standard integrated circuits are used such as 4000 series CMOS chips, and the designs are therefore obsolescence-proof as far as they can be. However some of the Appendices describe ancillary circuits developed some years ago which used special purpose devices which are now difficult or impossible to obtain. This material has been included to illustrate particular techniques which could still be implemented with some re-design to enable today's devices to be used. When this book was first written in 2001 most of the components required could be obtained from Maplin's and their stock codes are quoted where appropriate. The same has been done for other more specialist suppliers in a few instances.

Many novel design aspects are represented here. For the divider organ there are no cable looms between the tone generators and keyers, and the keyboards are speedily connected using IDC connectors. The free phase system actually needs fewer IC's than the divider one, and fully flexible tone forming is possible as the system does not depend only on sine waves.

In a volume such as this it is difficult to decide how to handle the large number of diagrams. Placing them close to the text which refers to them is an obvious choice at first sight, but in many cases a diagram is referred to in several places. Another method is to bind the diagrams separately, but this increases production costs. Consequently the diagrams for the main text have been placed at the end of the final chapter, and those for the Appendices have been attached at the end of each one.

No attempt has been made to include every possible technique. This is because I have dwelt only on those which are known to work, and in most cases this means I have used or assessed them myself. However the omission of other approaches does not necessarily imply they are of lesser value.

Great care has been taken to ensure freedom from errors, but it is probably expecting too much to assume that none have crept in. I should be grateful for my attention to be drawn to any errors or other problems, and email is the most convenient (cep@pykett.org.uk). Or use the address or phone number in the EOCS membership list, but when writing please enclose a stamped addressed envelope if a reply is required.

C E Pykett
January 2003

ABOUT THE AUTHOR

Colin Pykett trained as a physicist, completing his PhD research at King's College London in 1970. He is also a Fellow of the Institute of Physics. He retired recently as Chief Scientist in a high technology organisation after a career during which he spent many years working in acoustics and digital signal processing. His musical training began at the age of six with the piano, and subsequently it widened to include the oboe and the organ. Initially he studied the latter instrument with the late Russell Missin at St Mary's Nottingham in the 1960's, and subsequently he studied with others of similar stature while at university. He has played the organ, sung in choirs and trained them in many churches over some 35 years. Dr Pykett has undertaken research in organ topics for over 20 years, particularly in the mechanisms of tone production in organ pipes. More recently he has investigated responsive mechanical and electric actions for pipe organs from both an experimental and a theoretical standpoint. He also acts as an organ adviser. Many contributions have been made to the literature, and the author also maintains a website at www.pykett.org.uk.

Colin has been a member of the Electronic Organ Constructors' Society for some 25 years. Not only has he contributed to the Society's activities over this period, but he has learnt much from it.

CHAPTER 1

Analogue Organs

There have been two step changes in electronic organ technology in the last 25 years or so. The first was the appearance of integrated circuits which suddenly made the job of building an analogue organ much simpler than hitherto. Probably the best and last of these was the Texas Instruments TMS 3617, a chip which made its appearance in the 1980's but only lasted a year or two. It enabled an octave of notes at 6 footages to be obtained from a single chip with only a single clock input. The reason for the brief lifetime of chips like the 3617 was because of the second step change - the sudden shift from analogue to digital technology that occurred in commercial organs in the 1980's. At a stroke this put the electronic organ beyond the realm of home construction for most people because of the specialist knowledge and resources required in both hardware and software. But there was also a realisation that, as prices came down, perhaps the imperative to build an organ at home was no longer there. But with the fantastic growth in power of personal computers it is now possible for the enthusiast to re-enter the digital organ field, and this topic will be explored in volume two.

Analogue technology is still worthwhile if you want simple circuits that can be tweaked to your heart's content, and it is not true to say that good results cannot be obtained. Being able to change the sounds easily is important, because this is impossible or extremely difficult with some digital designs. Unfortunately many digital organs consist merely of a hotchpotch of sound samples copied from several sources. Consequently a common result is a lack of blend due to incompatible timbres and their scaling and regulation, even if the individual stops are satisfactory. We shall return to these issues in volume two. Up to date analogue circuits can also be used to give a new lease of life to an existing instrument. If one is sensitive to such matters they also give a feeling that one is dealing with a proper musical instrument: the provision of circuitry to replicate the many pipes in an organ can indeed be an *artistic analogue* of the craft of the organ builder, particularly when dealing with free phase. Few would contend that any electronic organ, analogue or digital, could be indistinguishable from a pipe organ because it is plainly not true. However this is also true of the humble reed organ which in its day also strove to imitate pipes, yet now it is regarded as an instrument in its own right. So it is with electronic organs if they are designed and built artistically.

Broadly speaking there are two sorts of analogue organs – divider and free phase (FP). In the simplest form of divider organ, a single oscillator is used to derive the 12 notes of the scale for the top octave of the organ, and the frequency of each of these is then successively divided by two to provide the lower octaves. A more or less complex keying system is required to enable multiple footages to be applied to tone forming filters, and the keying circuits also have to provide envelope shaping to prevent an unnaturally sudden attack and decay being applied to the sounds. Objections to the divider system, apart from the complexity of the keying arrangements, centre around the phase locked nature of the sounds in that the frequencies of notes an octave apart are exact multiples of two. However there are ways of mitigating this problem at some increase in complexity.

In free phase each note has its own oscillator which is independent of all others, so there is no phase locking as with the divider system. However a rank of FP oscillators can only provide one footage, unless complex keying circuits (comparable to or even worse than those of the divider system) are employed to turn the system into an extension organ in effect. Therefore multiple ranks of oscillators are generally required, and the main problem then becomes one of oscillator stability as well as the sheer number of oscillators. Also the oscillators have to provide waveforms with many harmonics such as sawtooths if effective tone forming is to be applied. This can be a problem because the most stable oscillators only generate sine waves directly. In the days of analogue organs extremely effective examples of both types were made both commercially and by individuals, and it is important to avoid being too wedded to one camp or the other in my view.

In this volume circuit details will be given for both types, the reason being that it is relatively easy to build the circuits without specialist knowledge and facilities. And, as already mentioned, there is no reason why excellent results should not be obtained. The main drawback is that rather a lot of circuitry is required if a full scale instrument is to be constructed from scratch. Therefore the articles will direct readers towards approaches of maximum simplicity in this regard, using some novel ideas as well as more traditional approaches. It might be best to wait until you have read about both types of organ because you could be surprised at how economical a FP organ can be in terms of IC package count.

Before going further it will be useful to set out a specimen stop list of a small organ that is musically useful yet not too demanding to build if one so desired. This is important, as it then defines the requirements for much of the technical information which follows. There would not be much point giving pages of data without at the same time giving some idea of the sort of instrument I had in mind. The stop list is as follows:

SWELL		**GREAT**		**PEDAL**	
Geigen Diapason	8	Open Diapason	8	Major Bass	16
Stopped Diapason	8	Claribel Flute	8	Sub Bass	16
Geigen Principal	4	Dulciana	8	Octave	8
Nason Flute	4	Principal	4	Bass Flute	8
Block Flute	2	Fifteenth	2	Trombone	16
Double Trumpet	16	Trumpet	8	Schalmei	4
Cornopean	8				
Clarinet	8	Swell to Great		Swell to Pedal	
Clarion	4			Great to Pedal	
Tremulant					

Some features of this design are:

1. There are no mixtures or mutations. This is because the electronic simplifications introduced are so great that one has to be realistic about this. Introducing merely a Mixture on the swell and a Twelfth on the great would virtually double the complexity of the keying and generating systems just to accommodate these two extra stops. However expanding the system to include mutations will be described for those who want them.

2. On the swell the most highly developed chorus in a musical sense is the reeds. A satisfying full swell effect will be obtained, and the Clarinet is a useful solo voice. Next are the flutes, which will produce a positive organ – like effect. The two diapasons are different in character from those on the great.

3. The main chorus on the great is the diapasons up to Fifteenth, which are bold and assertive. The Trumpet is voiced more assertively than a chorus reed would be, but nevertheless it will not drown out the rest of the stops. The Dulciana and Claribel Flute, besides being useful on their own, can be used to accompany the swell reeds used as solo stops.

4. On the pedal the main point is the introduction of an independent four foot reed to enable Chorale Preludes to be played while using the swell for the manual parts, which would preclude coupling the swell four foot reed to the pedals.

Anyone who wished to construct this instrument would become the owner of a small but high quality organ which would bear comparison with many commercial (digital) products today. It could of course be expanded if desired. Divider and free phase tone generating systems will be described which could be used to realise the specification above. Either could also be used to form the basis of an entertainment or theatre instrument, but the only tone filters to be described will be those corresponding to the classical specification given. The divider version of the instrument will have a much better technical performance than one constructed with devices such as the TMS 3617, which was a temperamental device with several disadvantages. One of its worst shortcomings was the appearance of audible sub octave outputs on some footages, and its background noise performance was only just acceptable.

I must make it clear that an organ with exactly the specification above has not been built as an entity, nor have all the circuits to be described later necessarily been used in a full scale instrument. However the tone filters and waveforms in this instrument are identical to those in a larger one which I still possess and which readers are welcome to play. This should enable a good impression to be gained of the sound of the organ before any decision is taken to build it.

All circuits have been properly tested. Therefore readers can proceed with sufficient confidence to use the circuits as a basis for their own work, although some previous electronics experience is desirable. This is because I shall not be describing the details of things like power supplies, the elimination of earth loops, etc. The availability of test gear such as an oscilloscope and a frequency meter would also be advantageous, though perhaps not essential. Those without a scope might note that economical interfaces to turn your PC into one can be purchased (e.g. from Maplin's).

I shall be happy to engage with any reader who encounters problems, which can then be reported in the form of updates or revisions if necessary. Please see the Foreword.

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CHAPTER 2

A Divider Organ – Top Octave Generators for the Swell Department

The block diagram of the swell department of the small organ mentioned in chapter 1 appears in Fig. 1. It uses two TOG's (top octave generators) which feed adjacent octaves so that the corresponding notes, e.g. C3 – C4 etc, are unlocked in terms of frequency. (In this book bottom C of each department is called C1, the octave above C2, etc). Each TOG is driven by a very stable LC tuneable oscillator - it seems more natural being able to tune a musical instrument, and this is why I like using tuneable oscillators rather than crystals, though there are other reasons as well. These particular oscillators also allow a frequency-modulation tremulant signal to be injected very easily. Taking TOG 2 first, this runs at 2 MHz and feeds twelve 9-bit dividers, each one dividing by the appropriate number to produce the corresponding note of the equally tempered (ET) scale. The notes so produced form the top octave at 2 foot pitch.

This particular set of 12 divisors has been very widely used in almost every LSI analogue organ chip ever made but restricting them to 9 bits (i.e. less than 512) means that only an approximate ET scale is generated. Probably the most obvious drawback is the appearance of two perfect fifths between D/A and D#/A#, whereas a true ET scale has no perfect intervals at all. To my mind one can get too hung up about this sort of thing, as it is only a short time before even a pipe organ develops such tuning irregularities after it has been carefully tuned. However, a good discussion of the arithmetic involved appeared in EOM 176 (January 2001) in Trevor Skeggs's article. (Only last year I played a pipe organ which was supposed to be tuned to some obscure Baroque temperament, yet it had perfect intervals in exactly these positions in the middle octave of a principal stop!).

Nevertheless, to avoid the same perfect intervals recurring in every octave the other generator (TOG 1) runs at an offset frequency a semitone higher at 2.12 MHz. Moreover, since it only needs to provide a top octave at 4 foot pitch, it can use more accurate 10-bit instead of 9-bit divisors. These only generate a single perfect fifth in each octave instead of two, and because of the semitone offset this interval appears between E and B. Taken together therefore, considerable diversity has been introduced into the generator system as far as adjacent octaves are concerned and this largely compensates for the worst aspect of dividers, which is the reinforcement of the same frequency defects in every octave. To summarise, octaves 1, 3 and 5 have two perfect fifths at D and D#, and octaves 2 and 4 have one at E. Also the tuning of these two groups is completely independent. This degree of diversity will largely disguise the fact that the instrument uses frequency division for normal playing purposes. The main criticisms of divider organs stem from the fact that, in their simplest form, the same approximations and compromises are repeated in all octaves. These criticisms arose from early instruments 50 years ago and have been impossible to shake off.

Each TOG feeds an array of standard divide-by-two circuits, which together provide enough square waves to cover the range from 16 foot C1 to 2 foot B5 (2 foot C6, the topmost pitch of the top note, breaks back to 4 foot). All of these square waves are then fed to the keying circuits.

The simple statement in the last sentence disguises enormous tedium and complexity. Feeding 72 wires from TOG 1 and its dividers and 96 from TOG 2 to the keying system illustrates a main drawback of analogue organs. Some might not have too much of a problem with it. After all, it is still a lot simpler than building a pipe organ, or even wiring many a pipe organ console. But if there is a way to avoid it, then it should be described. In fact there are several ways, some of which use computers. I shall describe a method which does not involve computers although it uses more integrated circuits in the keying system. So if you threw your hands up in horror at this point, please be patient a little longer. Also, if you panicked at the mention of only square waves, please be aware that they will be used to derive a variety of other waveforms via the method of staircasing. This is definitely not a square wave organ as far as tone forming is concerned!

Each key has a single contact which allows four square wave pitches (16, 8, 4, & 2) to be applied to the tone forming system via a gradual attack and decay circuit (soft switching). The contact activates probably the simplest form of electronic switch I have ever designed, yet the most effective. It has an on/off ratio of at least –70 dB and probably in excess of –80 dB. Thus breakthrough in the form of "beehive" is completely inaudible. In fact the performance of the keyers is so good that particular care will have to be taken with things like circuit layout, power supply design, earth line routing, etc if it is to be realised in practice.

The outputs of the keyers are divided into four groups or segments corresponding to the octaves of the keyboard (segment 4 includes both octaves 4 and 5). This enables each tone filter to be subdivided into the same four segments so that proper voicing and regulation can be carried out octave by octave. This is because the tone quality and volume of an organ stop both vary across the compass, largely due to the variation of pipe scale (the ratio of diameter to length) across the rank. Only the finest analogue organs used this technique, which makes a tremendous contribution to the fidelity of the sounds produced. Each segment of each filter derives an optimum waveform by picking off square waves as necessary from the four busses at 16, 8, 4 & 2 foot pitches, and weighting them. It is just as important to derive the correct waveforms as it is to design the filter networks themselves, rather than merely using one or two "stock" waveforms such as squares or sawtooths. In the stop list already described the Clarinet, for example, uses mainly the odd harmonics from a square wave at 8 foot pitch, but with low levels of even harmonics obtained from another at 4 foot pitch. In the diagram it is implied for simplicity that each tone filter only picks off a single square wave from each segment, but in practice more will usually be required for staircasing purposes.

The tone filters then feed one of two loudspeaker channels, to enable spatial separation and acoustic mixing to overcome the fact that the stop signals are drawn from the same generators. However an optional feature can be used to further compensate for this by applying a slowly varying phase shift to some filter outputs. The rationale underlying the distribution of stops to output channels and which of them need phase shifters is developed in chapter 6. Finally the signals in the two output channels are applied to a dual channel swell control which is tone-compensated to simulate the characteristics of a real swell box.

Power Supply

The circuits to be described all work at 15 volts unless otherwise specified. Lower voltages are not recommended for the following reasons:

1. Ordinary CMOS works fastest at its highest recommended voltage of 15V, and this is important in the TOG's where very high speed resetting of counters is required in order to realise the desired divisor values.

2. In the analogue circuitry it is important to allow sufficient headroom for situations when many notes are keyed at once. Otherwise distortion will occur when peak signal excursions approach the power rails.

3. It is desirable to use as high a voltage as practicable at the key contacts to overcome contact resistance due to contamination. In particular if silver wires are used, they will not work reliably under 6 volts owing to reactions with atmospheric gases such as sulphur dioxide.

4. The design of the keyers is such that they switch 15 volt square waves. This is much too high for use in subsequent stages, so the necessary attenuation also attenuates breakthrough signals when the switches are off. This design feature contributes around 20 dB to the high performance of this switch in terms of its on/off ratio.

A full circuit for the power supply will not be given but an outline schematic diagram appears in Figure 2. Separate supplies for the digital and analogue sections of the system are derived from a common transformer/rectifier and smoothing components, and separate analogue and digital ground lines are run. In this diagram and in the circuits which follow, digital power rails are identified by the use of D+15, and analogue ones by A+15 and A-15. The two ground lines are denoted by DGd and AGd. These precautions are intended to prevent the high performance of the keying circuits being degraded by residual breakthrough ("beehive") occurring via the power and ground lines. However, as many readers will have experienced, the most difficult and obscure problems can sometimes arise despite the most careful attention being paid to such matters. It might be found necessary to join the analogue and digital grounds in certain places, particularly near to the keying circuits. There are generally three ways to do this, and all should be tried to see which gives the best results:

1. The lines can simply be joined together

2. They can be joined via a low value resistor such as 10 to 100 ohms.

3. They can be joined via two diodes in parallel but with opposite polarities.

Heavy gauge wire should be used particularly for the ground lines between power supply and circuit boards, and 24/0.2mm equipment wire would not be overdoing it. Of course, this is not because of excessively high currents but because of the need to minimise line impedances and hence feedback of troublesome signals throughout the system.

Some circuit diagrams include specific decoupling components. However even if these are not shown, digital circuitry should be decoupled with a 100n ceramic capacitor close to every three or four chips. Boards containing analogue circuits should have 100 uF electrolytics placed close to where the power lines enter the assembly.

Finally, you should not attempt to build a power supply from scratch unless you are fully experienced and conversant with the health and safety implications of mains voltages.

Oscillators

Each TOG is driven by an LC tuneable oscillator for simplicity using the circuit of Fig.3. Crystal oscillators could be used but they are too accurate for this application. A crystal has an accuracy of typically 0.001% which is at least 10 times too precise for a musical instrument. If crystals at the specified frequencies were used to drive the two TOGs there would be no perceptual unlocking of adjacent octaves. It is possible to vary the tuning of a crystal oscillator using phase locked loops, but this introduces significant additional complexity when none is needed. Merely connecting a trimmer across the crystals would not allow them to be pulled in frequency sufficiently.

I have used the Figure 3 circuit for many years and it is extremely stable; one of my instruments used 21 of them and it scarcely ever needs tuning. In fact it only gets tuned when the back is unscrewed for other reasons every two years or so. It is recommended that the coil is wound on a ferrite core kit type RM 6/160, available from RS Components (stock number 228-214). It can be ordered direct from the RS website or via Electromail, and it has been on their stock list for at least 20 years. The coil only needs 8 + 8 turns of 27 SWG wire (0.4 mm diameter), although the gauge is not too critical. The main problem is to avoid losing the tiny components of the kit when you open the box! An alternative, cheaper, construction is to use a standard 0.25 inch diameter open plastic former accepting an ordinary ferrite core. The same wire gauge can be used but about 22 + 22 turns are now needed. In this case the windings will need to be tightly stabilised against movement, for example by wrapping them with adhesive tape and then over-winding with a layer of lacing cord. I have no real experience of long term stability for this latter type of coil, but have used a number of them successfully for various experimental purposes. The main drawback is that they cannot be mounted too close together or else they interact. (Having a closed magnetic circuit, the RM 6's can be mounted very close with no problems).

The tuning capacitor is polystyrene with a value of about 82 – 120 pF. It should be mounted on standoff pins to facilitate adjustments, and a value of 100 pF should be used initially. If the required frequency cannot be obtained by adjusting the slug, it should be replaced by a larger or smaller unit. If the open type of coil is used a somewhat higher value of capacitance is usually needed, about twice the above values.

The oscillator output is buffered by applying it to a 4093 Schmitt trigger chip using all four sections in parallel. Each output is then used to drive groups of 3 divider strings which generate the top octave. When measuring frequency the meter probe should be

applied to the output of a buffer section, not directly to the oscillator which otherwise would be loaded by the capacitance of the probe and would therefore change in frequency. If no frequency meter is available the frequency of the two oscillators can be set once their divider chains have been constructed. As these provide audio frequency outputs they can be tuned by ear against a tuning fork or by other means.
The layout of the oscillator circuit is not critical, and it works fine using ordinary stripboard.

Tremulant

The tremulant (Figure 4) is a low frequency phase shift oscillator which varies the bias on the two oscillator transistors. The modulating voltage from the tremulant is injected into both oscillators via a 22K resistor seen in Figure 3, and the 100n decoupling capacitors must not be omitted or the two oscillators may interact.

Speed and depth controls are as in the diagram. Because the tremulant circuit is connected directly to the oscillators it should be mounted fairly close to them, and long leads should not be connected to the on-off switch shown in the diagram. Therefore this switch should be a small relay or electronic switch of low impedance operated by the remote stop key or drawstop.

Top Octave Generators

Referring to Figure 5 (a) the dividers for each TOG consist of twelve 4040 12-stage binary counters, of which one is shown giving a divisor of 536. This corresponds to top B of the 4 foot top octave (TOG 1), whose frequency is 3953 Hz for a clock input of 2.119 MHz. Thus 3953 equals the clock frequency divided by 536. The clock inputs (pin 10) of each group of three counters are driven by one of the outputs of the 4093 trigger circuit which acts as a buffer between the clock oscillator and the dividers (see Figure 3).

Each divider is prematurely reset before it reaches its maximum count by a diode AND gate as shown. The diodes are connected to the outputs of the 4040 in a way which represents the binary equivalent of the desired divisor, and they cause the MR (reset) input at pin 11 to go high when all the selected outputs are simultaneously high. The divided output is taken from the pin which provides the best square wave in terms of its having a mark-space ratio closest to unity and in this case it is pin 12. For clarity the AND gate is redrawn in Figure 5 (b), where point X (and hence MR) goes high only when all the diode cathodes are high simultaneously. This only happens in this case when the count has reached 536. The use of slightly non-symmetrical square waves is of little consequence in the top octaves because their harmonics rapidly go beyond the limits of audibility, therefore the ear is not sensitive to harmonic content at these frequencies.

Diode gating is used purely for simplicity, to avoid many additional logic packages, but it is not always straightforward to get it to work reliably at the high clock frequencies used here. Many experiments have been carried out and it is believed the arrangement shown in Figure 5 is robust and should be reliable. Tests have been made on circuits used by other authors, including those of David Ryder in EOM 90 (April 1980). However the reliability of the circuits depends on factors such as the supply voltage and, equally importantly, on the make of chip so other people's arrangements might not always work here. I tested two types of chip, the TC4040BP (a sample several years old) and the HCF4040BE (recent). The circuit used here was reliable with both of these whereas some other circuits were not. The latter chip is of interest because it is that supplied by Maplin's at the time of writing. Nevertheless if problems are experienced try changing the value of R1, although care should be taken to ensure it does not go so low that the current sinking capability of your chip is exceeded. The value shown is about the lowest that can be used on a supply of 15 volts. R1 + R2 introduce a time stretch in the reset pulse in conjunction with the stray capacitance at the input of the reset pin, and R2 can be varied over a wide range if necessary. However, let me repeat that no problems have been experienced with the circuit shown. The TOG even works on 12 volts with almost no change in frequency, although this is not recommended for the reasons already mentioned.

The outputs of the 4040's are each buffered using one section of a hex buffer (4049 or 4050) to isolate them from succeeding circuitry which could present a significant capacitive load, such as the wiring to the keying circuits. This is particularly important for those outputs which are also connected to the reset AND gate (see tables below), as any degradation of the pulses here could affect the reliability of the reset circuit. Two buffer packages are needed for each group of twelve 4040's. When the circuits are built attention should be paid to keeping stray capacitances to a minimum near to the reset pins. This will be particularly important if stripboard is used, and tracks should be cut so that they are no longer than necessary. Again, though, the fact that the circuits can be made to work on breadboards without problems suggests that they are reasonably robust to strays. They are also entirely reliable on stripboard - the master oscillator, the 4093 clock buffer, the 12 TOG dividers and their hex buffers can all be contained comfortably on a single board 300 by 100mm with room to spare (Maplin stock code JP51F). Remember to decouple the logic packages copiously with 100n ceramics as previously recommended.

All of the AND gating details are in the tables below. Note that the frequencies shown are those actually generated, and they differ slightly from the exact frequencies of the equally tempered scale owing to the finite precision of the divisors.

Gating details for TOG 1 (Clock frequency = 2.119 MHz)

Note	o/p frequency (Hz)	Divisor	Diodes from pins	Output from pin
B	3953	536	3,5,14	12
A#	3731	568	2,3,5,14	12
A	3520	602	3,4,5,7,14	12
G#	3321	638	2,3,4,5,6,7,14	12
G	3135	676	2,6,13,14	12
F#	2959	716	4,5,6,13,14	12
F	2795	758	2,3,4,6,7,13,14	12
E	2636	804	2,6,12,14	14
D#	2490	851	3,4,7,9,12,14	14
D	2349	902	6,7,12,13,14	14
C#	2217	956	2,3,5,6,12,13,14	14
C	2094	1012	2,3,4,6,12,13,14	14

Gating details for TOG 2 (Clock frequency = 2.000 MHz)

Note	o/p frequency (Hz)	Divisor	Diodes from pins	Output from pin
B	7905	253	2,3,4,5,6,9,13	13
A#	7463	268	5,6,12	13
A	7042	284	3,5,6,12	13
G#	6644	301	2,5,6,9,12	13
G	6270	319	2,3,5,6,7,9,12	13
F#	5917	338	3,4,7,12	13
F	5587	358	2,4,6,7,12	13
E	5277	379	2,3,4,5,7,9,12	13
D#	4975	402	3,7,12,13	12
D	4695	426	2,5,7,12,13	12
C#	4435	451	4,7,9,12,13	12
C	4184	478	3,4,5,6,7,12,13	12

Now that TOG's are no longer available in the form of integrated circuits it was considered worthwhile to describe in some detail how to make them using standard CMOS packages. Even if you do not want to use them in a complete instrument they have other uses, of which one is the nucleus of an electronic tuning aid. Another is as a programmable frequency source for use in computer music experiments, as it is much easier to get a computer to select one of several ready-made frequencies than to program it to generate the waveforms from scratch.

Generating the lower frequencies

The output of each TOG buffer section, of which there are 24 in all, is sent to the keying circuits which require that frequency and it is also applied to the clock input of another counter which generates the remaining pitches. The block diagram (Figure 1) should make this clear. 4024 7-stage counters are probably the most convenient. Thus twenty four 4024's are required. Since these operate at audio frequencies the circuits are not critical, though again the chips should be decoupled at intervals by 100n ceramics to avoid spurious pulses propagating throughout the entire system.

However before building the 4024 divider arrays, pause to recall the earlier discussion about how to distribute the generator outputs to the keyers. The block diagram (Figure 1) implies that a brute-force approach is used in which bundles of multiple wires are used to interconnect the generators to the keyers in a matrix fashion. Anyone who has done this in the past will understand what is involved. But I mentioned that there are ways to reduce the effort involved at the expense of additional complexity elsewhere. One way will be mentioned shortly when the keying circuits are described, so it is recommended that you wait until that section has been digested before deciding on your preferred approach.

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CHAPTER 3

Swell Keying Circuits

The keying circuit for each note uses a 4016 quadruple analogue switch in an unusual way which enables a gradual attack and decay to be obtained, rather than the normal abrupt on/off action. The circuit for one note is shown in Figure 6. There is only one single pole normally-off contact per key. Pressing a key causes +15 volts from the analogue supply to be fed to the keying circuitry, and the key contact common line is connected to the power supply via a low value resistor. This is purely a precaution should a short circuit arise somewhere in the keyboard wiring, perhaps while adjusting the key contacts.

R1 and C result in a rising voltage ramp at the output of the emitter follower which is applied to all four "Z" connections to the switch. The ramp therefore appears at the "Y" connections also, but it is chopped in each case by the square wave applied to the respective enable ("E") terminal. The four square waves are obtained from the tone generator system and represent the 16, 8, 4 and 2 foot pitches for the note in question. When the key is released the output square waves ramp down from 15 volts to zero with a longer time constant of (R1+R2)C, giving a just-perceptible sustain which somewhat compensates for the acoustic deadness of the average domestic room. Thus we have achieved a soft switching action for all four footage signals using only a cheap IC and one transistor per note. It is preferable to use the 4016 rather than the higher performance 4066 because the latter introduces larger switching spikes into the output when the switches are off. Although it might be possible to obtain 4016's at a lower price than 4066's, do not be tempted to use anything other than branded devices from a reputable supplier in this critical application.

A typical value of the time constant capacitor C is shown in Figure 6 but it varies across the keyboard as in the table below.

Octave	Notes	C
1	C1 – B1	4u7
2	C2 – B2	2u2
3	C3 – F3 F#3 – B3	2u2 1u0
4	C4 – B4	1u0
5	C5 – C6	1u0

Note the disposition of supply and ground rails in Figure 6. The analogue and digital grounds are shown connected by a dotted line, and there are several ways to interconnect these as discussed in chapter 1. The chosen method should be that which gives the lowest audible breakthrough ("beehive") when all keys are up. The breakthrough performance of this circuit is extremely good because, in the off state, there is no input signal to the switch at the Z inputs so there cannot be any breakthrough in the normal sense. Any which exists arises either as breakthrough from the enable inputs, or from strays in the circuit layout. Breakthrough from the enable inputs is not specified by all manufacturers, but because it is in the form of very short spikes there is little audio frequency content anyway. The 4016 is better in this respect than the 4066 as previously mentioned. Attempts to measure breakthrough were frustrated because of its low level, but comparative listening tests suggested it was well below 60 dB. This is further attenuated in the manner to be explained shortly. Breakthrough due to strays will usually be the dominant factor, and the power supply and grounding arrangements should be adhered to. Also the input signals from the tone generators should be kept away from the output pins.

The output signals from all the 4016's in an octave are routed to virtual earth mixers. A glance back at the block diagram (Figure 1) shows that there are four of these per segment, one per pitch. Segmenting the keyer outputs in this way enables signals from each octave to be fed to separate sections of the tone filters for better voicing and regulation to be achieved. (The outputs of the mixers for the top octave, octave 5, and the one below it are combined into a single segment as described presently).

It is convenient to use quad operational amplifiers for the mixers, one per octave, and the TL074 is recommended. These are high performance op amps, much better than 741's in terms of noise, input impedance and slew rate. The latter is particularly important because slew rate limited amplifiers introduce intermodulation distortion at high frequencies. Fortunately the TL074 is cheaper than the LM348, a quad version of the 741, which implies that most designers agree with the above remarks! If layout dictates that single op amps are more convenient then TL071's should be used, though this makes for a slightly more expensive solution but with more flexibility in layout. The mixer circuits for each octave are in Figure 7 (segment 4 is slightly different and will be dealt with shortly). Thus the twelve outputs of the same footage from each 4016 in an octave feed resistors connected to the summing node of a mixer, which then feeds the tone filters for that footage and segment. Resistor values are all 15k except for the top octave where they are 10k. The reason for this will be explained shortly. The feedback resistor is 1k5, therefore the mixers attenuate rather than amplify.

For an input resistor of 15k the attenuation is a factor of 10 or 20dB, therefore the breakthrough signals from the 4016's are also attenuated by 20dB which gives the overall switch configuration its high performance. With ordinary equipment it is impossible to measure the on/off ratio of this arrangement accurately but it is at least 80 dB and probably better. Any breakthrough which does occur is almost certainly due to strays and can be controlled by attention to power lines, decoupling and layout in the manner already described. The attenuation also means that the 15 volt square waves from the 4016's are reduced to about 1.5 volts, which allows adequate headroom in subsequent circuitry when several notes are keyed simultaneously.

As previously mentioned the outputs of octaves 4 and 5 (the top octave) are combined into one segment, segment 4. This is shown in Figure 8 for one footage, so this circuit has to be repeated four times as in the previous case. The combining is done at the mixers for octave 4, which each receive a unity gain input from the octave 5 mixers. This configuration allows the tone filters to have only four rather than five segments for economy, because applying tone forming to the top octave separately would be rather a waste of time. This is because the ear becomes insensitive to tone colour at higher frequencies. But note from Figure 7 that the mixer resistors for octave 5 are lower than those of the other octaves to boost the signal levels in this region by about 3.5dB, which compensates for the low pass effect of most tone filters. Combining the signals from octaves 4 and 5 in this way is simple but it introduces an extra phase inversion for the top octave signals. This has no audible consequence because the master oscillator from which they are derived is independent of that for octave 4.

An alternative way of combining the outputs from octaves 4 and 5 is simply to feed the 10k resistors of octave 5 directly into the octave 4 mixers. Octave 5 then has no mixer of its own. This scheme was adopted for the free phase organ yet to be described and illustrated in Figure 34, although the resistor values are different for the free phase case. However care must be taken to keep leads which are connected to the mixer summing nodes (inverting inputs) short or to screen them.

The values of the mixer input resistors (10 – 15k) have been chosen deliberately after much design work. They need to be relatively low to prevent the mixers picking up stray signals and thereby degrading the effectiveness of the keying circuits. But they then present a significant current load to the keying circuits, which is why an emitter follower has to be used to feed the 4016's. The source impedance seen by the mixer input resistors has to be low compared to their own values, otherwise cross-modulation will occur between the various footages. (This was a problem with the TMS 3617 which had high cross-modulation, resulting in audible suboctave outputs). Measurements on the present circuit showed a cross-modulation level between one footage and its suboctave to be better than –60dB, which is inaudible.

Reducing interconnections

Mention has been made already of means whereby the interconnection problem between the generators and keying circuits can be reduced. In the circuits just described it was assumed that each keyer (the 4016 chip) picked off its four square waves from the large wiring matrix suggested by the block diagram in Figure 1.

This problem can be reduced as follows. Instead of using the 7 by 12 and 5 by 12 divider arrays depicted in Figure 1, it is possible to incorporate the necessary frequency dividers with each keying circuit. The divider arrays can then be dispensed with, leaving only two sets of 12 wires from the TOG's which go to the appropriate octaves. This is sketched in Figure 9 and it is obviously a much simpler inter-wiring problem, but the keying circuits are now more complex.

The schematic diagram of each modified keyer is in Figure 10. One of the 24 reference frequencies from the TOG's (at 2 foot pitch) is applied to the clock input of a 4520 dual counter. A jumper selects which output is used to clock the other half of the counter, which then generates the four footages required by the keyer. Thus the jumper defines the octave in which the keyer is required to operate. Note that the output from TOG 1 is already an octave lower than TOG 2, so the jumper settings are in the table below. Note also that jumper settings b and d are redundant for the swell department but they will be needed for the great.

Octave	Jumper	Ref freq source
5 (top)	a	TOG 2
4	a	TOG 1
3	c	TOG 2
2	c	TOG 1
1	e	TOG 2

The penalty paid for simplifying the interconnection problem is an additional IC package per keyer. However there is no longer any need for the 24 packages making up the divider arrays in Figure 1, so only 37 extra packages are required. This will usually represent an attractive solution. The full topological wiring diagram for the revised keying system is in Figure 11, which shows two circuits. It looks worse than it is, and this diagram will be discussed more fully when constructional details are covered.

Additional pitches

The basic organ being described has only four pitches per note, hence no mutation pitches and therefore no mixtures either. It should now be possible to see how to expand the system to include additional pitches if desired and if you can face the additional complexity. By including an extra 4016 in each keying circuit 8 pitches per note could be derived. However there would also be a need to bring in extra reference (quint) pitches to each keyer, and to provide an additional 4520. Extra mixers would also be required at the outputs. In other words the keying system doubles in size, measured in numbers of packages.

The complexity of such a system gets rather mind-boggling in my view, and that is why this description is of only a small organ with a deliberately constrained stop list.

Constructional Notes

A plan view of a suggested constructional scheme is sketched in Figure 12. It consists of a mother board/daughter board arrangement, with each mother board (MB1 – MB3) connected to a back plane. The back plane is divided into two unequal parts. The need for tedious hand wired cable looms has been completely eliminated, and it is believed this is the first published design for an analogue organ using only standard components with this major advantage. Even the wiring to the key contacts has been made simple by using ribbon cable and IDC headers which plug into the mother boards.

With the exception of the top note board, each daughter board contains two keying circuits as in Figure 11. The daughter boards are attached to a mother board in groups of six, making up one octave of notes (a seventh board is necessary in the top octave if top C is required). Each mother board accommodates two octaves of keyers, except for the top octave. The mother boards also contain the TL074 mixer amplifiers for each octave and the connections to the key contacts. The three motherboards are attached to the back planes which supply the reference frequencies and power. Also attached to the back planes are the two TOG's which are placed at opposite ends of the assembly so that they are close to the octaves which they supply. The tuning coils on each TOG board are arranged so they can be adjusted easily.

The circuits can either be built up on strip board, or PCB's could be designed without difficulty. If strip board is used, 30 by 10 cm boards are recommended with 39 tracks across the smaller dimension (Maplin stock code JP51F). Connections between the boards are made using 32 way plug and socket strips which are cut as necessary (Maplin DC17T and DC18U).

The dual keyer circuit was introduced earlier (Figure 11). This layout maps directly to strip board and it could also be used as the basis of PCB artwork, but note that it is not drawn accurately to scale. Layout is important for this board because of the need to prevent breakthrough between tracks carrying 15 volt square waves and the four virtual earth output lines leading to the footage mixers (which are on the motherboard). Stray capacitance must be kept below about 5 picofarads between any signal and output track. This is not so difficult as it might appear because strip board has a capacitance between adjacent tracks of about 1 pF per 2.5 centimetres, thus only the most careless layout would result in the strays getting out of hand. Consequently it is only necessary to keep the output tracks well away from those carrying signals, and to arrange guard tracks either side of the four output tracks. In this layout the guard tracks are the A+15 and D+15 lines which are at ground potential as far as breakthrough signals are concerned. If the layout is changed these aspects must receive close attention if beehive is to remain inaudible. If a PCB is designed it is worth considering the use of ground plane techniques as used in RF applications, in which (grounded) copper is allowed to remain over all unused areas on both sides of the board. There is no danger of breakthrough between the four output tracks themselves because these are current summing nodes, so the voltages appearing on them are minute. Note the 100n decoupling capacitors for the D+15 and A+15 lines.

The two relevant reference frequencies for each daughter board are picked off from the twelve lines running across the motherboard (see later for motherboard details). As only two frequencies per board are required (only one for the top note) it is only necessary to use two DC17T/DC18U connector pairs in the appropriate positions. This also prevents any unused frequencies getting onto the board at all as a further precaution against breakthrough. The maximum frequency (note B) is kept at one margin of the board to minimise breakthrough, which of course increases with frequency. The two lines labelled ck1 and ck2 in the diagram are connected by wire links to the two frequency inputs, and the 10 megohm resistors protect the chips from static build-up if the board is handled while disconnected from the motherboard.

A row of 8 connectors is used for the four footage outputs to the summing nodes (virtual earth points) of the mixers and the four power lines, and they are physically separated from the nearest frequency input (note C) with breakthrough considerations again in mind. The key contact inputs are at the other edge of the board, and although all 12 or 13 notes come in to the motherboard it is only necessary to provide two connectors for the notes actually used by the board (only one for the top note). These are joined by wire links to the lines marked K1 and K2.

The board is set up to provide its notes in the correct octave by setting the two jumpers appropriately. In fact these are again wire links. Note also the provision for interconnecting the analogue and digital grounds referred to earlier.

Top note

Providing the top note (C6) is generally rather awkward for one reason or another, and there are arguments in favour of omitting a note which is scarcely ever played. After all, a good many respectable pipe organs have less than 61 notes! However in this design it has been catered for by a 13 note octave architecture, and by allowing for a separate top note daughter board which only has one rather than two keying circuits

There is no 2 foot reference frequency for the top note because the highest one available from TOG 1 is for top B (B5). Therefore the reference frequency for the octave below, C5, has to be used but in turn this means that the arrangement of the 10k summing resistors into the mixers is different, as shown in Figure 13. Thus the 2 foot pitch breaks back to 4 foot for C6 only, whereas the other pitches for the top note are unchanged.

Motherboards

The motherboard arrangement is sketched in Figure 14, which is a view of the side of the board containing both the components and the daughter boards standing off at right angles. Thus the tracks, which are shown for purposes of explanation, would normally be on the other side of the board. All connectors shown in the diagram by the blobs are Maplin type DC18U (plug strips) and DC17T (sockets). Sockets to accept the daughter board plugs are mounted on the motherboard in the positions shown by the blobs. But at the left hand end of the board plugs are used to mate with sockets on the back planes. This accords with standard practice, in which protruding pins are never used to inject signals or power from one board to another; sockets should be used in these positions. A single strip board of size 30 by 10 cm with 39 tracks is a simple and convenient way to construct the motherboards (Maplin type JP51F).

At the top left are the 12 input plugs from the appropriate TOG, the highest frequency representing the note B using the topmost track. These 12 tracks run the whole length of the board, and each daughter board picks off the two frequencies for which it contains the keying circuits (shown by the blobs). As previously explained there are two octaves of keyers per motherboard, except for the top octave, and the octaves are grouped so they are nearest to their respective TOG as in Figure 12 .

Below the TOG inputs are the power line plugs which also run the whole length of the board, except for the A-15 line which feeds only the two TL074's – it is not needed by the daughter boards. At the point where the power lines enter the board there are some decoupling components.

Each octave has its own group of 13 key contact tracks, and again each daughter board only needs to pick off its two relevant lines. Note that these lines do not run the whole length of the motherboard. The keying lines for each octave are connected to a 16 pin DIL socket which accepts an IDC header (Maplin type JH36P) with a 16 way ribbon cable (Maplin type XR73Q).

The four footage outputs from each octave of daughter boards occupy four tracks which feed the TL074 mixers and their associated components, all mounted on the motherboard close to the respective octave.

More detail of the connections and suggested component placements is in Figure 15, and to conserve board space all resistors in this diagram are mounted vertically with the wire at the top bent back towards the board. As well as all the matters just discussed, this diagram also shows the 5 way output connectors which convey the four footages to the tone filter assembly via flying leads (except for octave 5, whose lead goes into the octave 4 mixer). PCB latching headers and sockets are suggested here. The units on the motherboard are Maplin type FY93B and the flying leads are connected to crimp terminals type YW25C which fit into housings type BH66W. These terminals can also be soldered instead of crimped. For octave 4 only an additional connector is used to receive the outputs from octave 5, and that on the motherboard is formed from DC18U plugs. The outputs from octave 5 are on the flying leads just mentioned, and these need to terminate on a small piece of strip board containing four 1k5 resistors (see Figure 8 to clarify the circuit) which then connect to DC17T sockets. The arrangement is sketched in Figure 16.

The 16 pin DIL sockets for connecting to the key contacts are visible in this diagram, one per octave, and depending on the keying and coupling arrangements which will be discussed subsequently, the other end of the ribbon cable might be connected directly to its respective octave of key contacts. Alternatively it might connect to a further assembly containing the couplers, a multiplex keying system, etc. Note that the 15 volts keying voltage appears at pin 14 of the DIL socket for connection to the key contact common line for that octave.

The keyboard wiring to the 16 pin DIL sockets as sketched in Figure 15 would actually result in connections to adjacent keys not occupying adjacent wires in the 16 way ribbon cable. This is because of the way the ribbon cable mates with the DIL header. This should present no problem in principle because the wires at the remote end of the ribbon cable can simply be fanned out to the correct key contacts. However if it is desired to adjust this, the wires in the ribbon cable connect to the pins of the DIL socket as follows:

Pin 16 connects to one extremity of the cable (right hand or left hand depending which way you are looking at it). The next (second) wire then connects to pin1, the third to pin 15, the fourth to pin 2, etc, etc.

Recall that in Figure 6 a low value resistor (33 R) is shown in series with the keying common line to prevent short circuiting the power supply accidentally when adjusting the key contacts. It is not essential to include this, but a convenient place for the resistor is to put it in series with the keying common line connected to pin 14 of the DIL socket in Figure 15.

Back planes

A front elevation of the divided back plane is in Figure 17. This diagram will become clearer if it is related to the plan view already shown in Figure 12. Thus we are now looking at the face of the back planes from which the motherboards project at right angles, and therefore these come towards us out of the plane of the diagram. The daughter boards are not depicted but they run parallel to the back planes in the direction shown by the arrows. As in the previous cases the track runs are shown, but in reality they would normally be on the back face of the back planes.

The two back planes each contain a TOG and its associated keying circuits, and there is no electrical connection between the two planes apart from common power supplies. The same Maplin type JP51F strip board can be used to construct the back planes, which only contain power and reference frequency tracks. The power tracks are doubled up to reduce impedances, apart from that for the A-15 supply. The duplicate tracks are interconnected at intervals, as shown in more detail in the expanded diagram. DC17T sockets are inserted in the back planes in the positions shown by the blobs, and these connect with DC18U plugs on the motherboards. (The power supply connections could also use these connectors but a more robust solution would be preferable, such as YW12N headers on the back planes and BH65V cable shells connected to the power leads).

CHAPTER 4

Top Octave Generators & Keying for the Great and Pedal Departments

Great

The great department, shown in Figure 18, uses an independent pair of TOGs for several reasons. Firstly, if the swell TOGs were used then both keyboards would be affected by the tremulant. Secondly, the four footages on the great are at 8, 4, 2 & 1 foot pitch whereas those on the swell are 16, 8, 4 & 2 foot. Thirdly, the more TOG's the merrier in a divider system, where it is essential to get as far away as reasonably possible from the crude concept of a single set of generators feeding the whole organ. However it is not unreasonable to economise by using the great TOGs for the pedal department as well, and this is indicated in Figure 18.

Because of the need to generate pitches an octave higher for the great, it is not possible in this case to use 10 bit divisors for TOG 1. Thus both TOGs use 9 bit divisors but their clock frequencies are offset by a semitone as for the swell and for the same reasons. This means that, while a pair of perfect fifths occur in every octave, their positions are shifted in adjacent octaves. But the main reason for using two TOGs remains as before, which is to unlock the tuning of adjacent octaves.

Construction of the oscillators and the use of 4040 dividers with diode gating is the same as for the swell, and diode gating details are in the tables below. Those for TOG 2 are identical to TOG 2 for the swell, but TOG 1 is different in this case. The frequencies shown are those actually generated, and they differ slightly from the exact frequencies of the equally tempered scale owing to the finite accuracy of the divisors.

Gating details for TOG 1 (Clock frequency = 2.119 MHz)

Note	o/p frequency (Hz)	Divisor	Diodes from pins	Output from pin
B	7906	268	5,6,12	13
A#	7461	284	3,5,6,12	13
A	7040	301	2,5,6,9,12	13
G#	6642	319	2,3,5,6,7,9,12	13
G	6269	338	3,4,7,12	13
F#	5919	358	2,4,6,7,12	13
F	5591	379	2,3,4,5,7,9,12	13
E	5271	402	3,7,12,13	12
D#	4974	426	2,5,7,12,13	12
D	4698	451	4,7,9,12,13	12
C#	4433	478	3,4,5,6,7,12,13	12
C	4188	506	2,3,4,5,7,12,13	12

Gating details for TOG 2 (Clock frequency = 2.000 MHz)

Note	o/p frequency (Hz)	Divisor	Diodes from pins	Output from pin
B	7905	253	2,3,4,5,6,9,13	13
A#	7463	268	5,6,12	13
A	7042	284	3,5,6,12	13
G#	6644	301	2,5,6,9,12	13
G	6270	319	2,3,5,6,7,9,12	13
F#	5917	338	3,4,7,12	13
F	5587	358	2,4,6,7,12	13
E	5277	379	2,3,4,5,7,9,12	13
D#	4975	402	3,7,12,13	12
D	4695	426	2,5,7,12,13	12
C#	4435	451	4,7,9,12,13	12
C	4184	478	3,4,5,6,7,12,13	12

The great keying system is similar to the swell except for a few details. Firstly all references to 16, 8, 4 and 2 foot pitches in the previous chapter should now be read as 8, 4, 2 and 1. Secondly the jumper settings for the daughter boards are different, as shown in the table below.

Octave	Jumper	Ref freq source
5 (top)	a	TOG 2
4	a	TOG 1
3	b	TOG 2
2	c	TOG 1
1	d	TOG 2

Thirdly a top octave (C5 – B5) at 1 foot pitch is not generated, so this has to break back to 2 foot. The changes to the mixer connections on the corresponding daughter boards are shown in Figure 19. For the same reason the 4, 2 and 1 foot pitches for the top note (C6) are all the same frequency, as shown in Figure 20.

A further optional matter might be to use different attack/decay capacitor values in the keying circuits. This is because the pitches available on the great are an octave higher than on the swell, so a correspondingly faster attack might be to your taste. One suggestion is to shift the swell capacitors down an octave, thus:

Octave	Notes	C
1	C1 – B1	2u2
2	C2 – B2 F#2– B2	2u2 1u0
3	C3 – B3	1u0
4	C4 – B4	1u0
5	C5 – C6	470n

Pedal

A block diagram for the pedal department is in Figure 21. It uses the great TOGs but they are wired in a complementary fashion to the octaves. This means that when the pedals are coupled to the great, corresponding notes will be unlocked. Square waves at 16, 8, 4 and 2 foot pitches are derived from the keyers, and these are fed to 2 – segment tone filters. The first segment corresponds to the lowest octave and the second to the remainder of the compass.

No phase shifters are shown, as they would add little at the bottom end of the compass particularly for the lower pitch stops.

The keying system is similar to that for the great except that 16, 8, 4 and 2 foot pitches are generated rather than 8, 4, 2 and 1 foot. This is achieved by wiring the jumpers on the daughter boards as in the table below.

Octave	Jumper	Ref freq source
3 (top)	c	TOG 1
2	d	TOG 2
1	e	TOG 1

A possible arrangement of a combined great and pedal keying system is sketched in Figure 22. Note that the top daughter board for octave 3 of the pedal is only required for a 32 note compass. The footage outputs of octaves 2 and 3 of the pedal are combined on the motherboard for octave 2 in the same way as for octaves 4 and 5 of the swell and great.

As with the great, you might also wish to vary the keying capacitors as the predominant pitch on the pedals is 16 feet rather than 8. In this case the capacitors would need to be made larger, but the attack cannot be made too slow otherwise the 4 foot reed when used as a solo would sound unconvincing.

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CHAPTER 5

Tone Forming Filters

The tone forming filters for the divider organ will be described in this chapter. With little modification they can also be used with the free phase generators to be described subsequently. The filters and the waveforms they use have been designed by computer from the sounds of real organ pipes, which were digitised and subsequently manipulated by software. The basic design method was developed many years ago and it was described in *Wireless World*, October and December 1980. These articles were reprinted in EOM 121 (June 1986) and 123 (November 1986). However most of the filters to be described were designed quite recently to update various analogue organs which are still going strong. Therefore these designs supersede any which have been previously published, even in cases where the stop names are the same.

General Arrangement

The general arrangement of each tone filter is shown in Figure 23. As already outlined each filter consists of four tone networks, each one optimised for a segment of the keyboard. This enables tone quality to be varied across the compass as it is for real ranks of pipes. Each network is fed by a staircasing amplifier whose input resistors (the staircasing resistors) are designated SR1 – SR3 in the diagram, connected in this example to the 16, 8 and 4 foot square wave busses of the swell department. The busses carry square waves with an amplitude of 1.5 volts per note, supplied by the mixers in the keying system. The arrangements of the staircasing resistors is purely illustrative, as each filter segment has a unique set of resistors depending on the type of waveform it needs and the pitch of the stop in question. Thus the stop shown would be a 16 foot stop, and if it needed a "standard" staircase wave (with harmonic proportions similar to those of a sawtooth) this would be derived by staircasing resistors having values in the ratios SR1: SR2: SR3 = 1:2:4 approximately (e.g. 22k, 47k and 82k). However several stops require non-standard values of staircasing resistors where a specific harmonic recipe is needed in the driving waveform.

The staircasing resistors feed into the summing node of an operational amplifier via a capacitor C. This is a DC blocking capacitor with a value chosen to prevent audible thump when keying notes in the associated segment, a necessary precaution especially at higher frequencies because the keying circuits are not DC balanced. Generally the value of this capacitor is chosen so that it introduces about 1 dB of attenuation at the lowest frequency it has to handle. The capacitance is given by C= 311000 / f R, where C is in nF and R in kohm. R is the resistance of all the input resistors in parallel and f is the frequency of the lowest note of the segment in Hz. However you do not need to calculate this yourself as the values will be tabulated for each tone filter in due course. The blocking capacitors can conveniently be 35 volt tantalum electrolytics, with the negative side connected to the input resistors.

RFB is the feedback resistor for the staircasing amplifier, and a common value of 22k is used for all filters. This gives an amplitude into the filter networks of about 1.5 volts for a square wave, and about 2.5 volts for a "sawtooth" derived from three staircased square waves, in all cases symmetrical about ground. Therefore there is ample voltage headroom when multiple notes are keyed. The four staircasing amplifiers can conveniently be a quadruple unit, preferably the TL074, operating on plus and minus 15 volts.

The staircasing amplifiers each feed a tone network optimised for the associated segment, and the outputs of all four are then summed via regulating resistors. The summing amplifier is a TL071, and its gain is variable to allow the volume of each stop to be adjusted. The amplifier output is fed via a stop switch, and some form of soft switch is recommended to prevent clicks appearing when the stops are manipulated while notes are sounding. An effective soft switch is shown in Fig 24. The circuit was fully described in EOM 165 page 7 (February 1998), reproduced here at Appendix 5, and it is recommended you peruse this first. The –5 volt supply can be derived from a single 100 mA regulator (itself fed from the A-15 line) feeding all the switches via a common supply line.

Each filter unit can be built on a single card which plugs into a motherboard carrying the power lines (A+15, AGd, A-15 and -5), the signal outputs and the 16 square wave busses (8 for the pedal). One motherboard is suggested for each department (swell, great or pedal), and they can be constructed using Maplin type JP51F stripboards size 30 by 10 cm with 39 tracks. The tone cards can use Maplin PCB sockets type YW30H, with corresponding plugs on the motherboard (Maplin type YW14Q). Note that we are diverting from convention here in allowing power and signals from the motherboards to be carried on projecting pins rather than sockets, but this solution is so cheap compared to edge connectors for example that it is worth considering.

There are several types of filter network, all described later and illustrated in Figures 25, 26 and 27. However for the purposes of designing PCB’s all those in Figures 25 and 27 could use a common layout, with wire links inserted in the series signal path where particular components are not needed for a given stop. Unwanted components in the shunt paths would simply be omitted. Also a common layout would suffice for the circuits in Figure 26 a and b, with a separate layout for Figure 26 c. Thus only three layouts are needed for all the many variations of filter network design described later in this chapter.

The tonal effect of the organ will be strongly dependent on the loudspeakers used and the room in which it resides. Therefore the values of the regulating resistors may need to be adjusted if the breaks between the segments become audible, or if you feel the need for treble or bass shading different to that given. However a few stops have been designed with audible breaks in order to get the necessary strong variation in tone quality across the compass. These stops would not normally be used by themselves, so the breaks will in practice be disguised by the other stops used with them. If you indulge in on-site regulation, a useful tip to bear in mind is that a pronounced change in timbre between adjacent segments (i.e. a break) can be completely disguised by adjusting (regulating) the corresponding levels, sometimes by a remarkable amount. This is a most useful property of the human auditory system which works in our favour in this application.

Swell Tone Filter Networks

There are three distinct choruses on the swell, consisting of the two geigens, the three flutes and the reeds (not including the Clarinet which is better regarded as a solo stop). Because of the limitations on the keying system which restrict it to delivering four pitches per note, it was not possible to add more diapason upperwork. The only way to do it would have been to add a makeshift Fifteenth using only a single square wave, and this would not have been satisfactory. It would produce only an artificial hollow buzz, particularly in the bass, which would be too reminiscent of the cheap and nasty electronic organs of yesteryear.

On the other hand a single square wave is an excellent waveform for a 2 foot flute and this is made available here. It complements the other two flutes to produce a sprightly and attractive flute chorus.

The three chorus reeds give an excellent full swell effect even though there is no mixture. This is partly because these stops have been designed with this factor in mind, and the characteristics of the tone networks draw out harmonics from the waveforms which a mixture would otherwise have provided.

With this degree of tonal variety available from only nine stops, the tonal combination possibilities are considerable.

Geigen Diapason 8'

Considerable care has been taken with the diapasons on this organ. This one has a quite different character from the Open Diapason on the great, so that (unlike on many electronic instruments) the difference between the various diapasons is not merely in loudness. The staircasing resistors produce a waveform with higher 2nd harmonic levels than a standard staircase wave (i.e. a sawtooth equivalent), as in the table below:

Segment	SR1 (8')	SR2 (4')	SR3 (2')	C (nF)
1	22k	22k	82k	470
2	22k	22k	82k	220
3	22k	22k	82k	150
4	22k	33k	82k	47

(RFB = 22k for all segments)

Blocking capacitor values (C) are less critical in the bass segments than the treble, because the keying time constants are longer for the lower notes, producing less thump to start with. Nevertheless it is advisable not to use excessively large values.

The tone network circuit for each segment is in Figure 25a. Segments 1 – 3 have a 3rd order passive RC filter (R1 – R3 and C1 – C3), driven from the output of the corresponding staircasing amplifier. Segment 4 only has two sections. R4 is the regulating resistor which feeds into the inverting input of the TL071 in Figure 23. The level preset for this amplifier is 100k. Component values for the networks are as below:

Segment	R1	R2	R3	R4	C1	C2	C3
1	1k3	10k	47k	180k	470n	47n	10n
2	2k	15k	51k	270k	220n	22n	4n7
3	2k2	18k	100k	430k	68n	10n	1n0
4	7k5	100k	-	1M0	10n	1n0	-

Note the dramatic increase in the values of the regulating resistors (the R4's) as the frequency increases. This is a characteristic feature of all the tone filters for flue stops (less so for the reeds), and it reflects the lower sensitivity of the ear at lower frequencies (the frequency of maximum aural sensitivity is about 3 kHz). These particular values also reflect the characteristics of my amplifiers, loudspeakers and the listening room, hence the possible need to adjust the R4's to suit your environment as mentioned already.

Stopped Diapason 8'

The Stopped Diapason has been designed to be used either on its own or as a versatile chorus stop. In this mode it can form the foundation of an 8/4/2 foot flute chorus, or it can be used with the swell diapasons. It is a genuine Stopped Diapason rather than a sweeter toned lieblich type of stop, and it has an attractive reedy character in the upper part of the compass just like the pipework from which it was derived.

The stop needs mainly odd harmonics but with low levels of evens as well. These are derived as in the table below:

Segment	SR1 (8')	SR2 (4')	C (nF)
1	22k	180k	220
2	22k	180k	150
3	22k	180k	68
4	22k	180k	33

(RFB = 22k for all segments)

The tone network is in Figure 25b, but not all segments use all the components. This is made clear in the following table:

Segment	Cin	R1	R2	R3	R4	C1	C2	C3
1	-	3k6	18k	82k	390k	470n	47n	10n
2	-	2k0	9k1	68k	1M0	470n	47n	4n7
3	-	2k0	9k1	-	2M2	220n	22n	-
4	100n	2k0	9k1	-	2M7	100n	4n7	-

(Level preset 220k)

Geigen Principal 4'

This stop is different in several ways to the Geigen Diapason, to reflect the different scaling and voicing of the corresponding organ pipe ranks. The waveform does not have an augmented second harmonic – it has the same harmonic proportions as a sawtooth wave. Also the corner frequencies of the filter networks are different. Therefore when used together the two stops combine to produce an effect very similar to what one would expect with two properly proportioned ranks of pipes with different scales and voicing.

The waveform components are:

Segment	SR1 (4')	SR2 (2')	C (nF)
1	22k	47k	220
2	22k	47k	100
3	22k	47k	47
4	22k	47k	22

(RFB = 22k for all segments)

The filter networks use the circuit in Figure 25c. R3 is the regulating resistor.

Segment	R1	R2	R3	C1	C2
1	1k8	18k	330k	100n	10n
2	1k3	13k	390k	100n	10n
3	910R	9k1	360k	100n	10n
4	910R	9k1	470k	68n	6n8

(Level preset 100k)

Nason Flute 4'

The waveforms and filters for this stop are different to those for the 8 foot Stopped Diapason. Thus the stop can be played on its own an octave lower to get a gentler sort of flute, and the two together are also most effective. It also combines well with the Geigen Diapason. When used with the Clarinet a quite different type of solo tone quality is imparted. Swell 4 foot flutes are relatively uncommon, and this one significantly augments the tonal capabilities of the department.

The waveforms are pure square waves at 4 foot pitch for all segments:

Segment	SR1 (4')	C (nF)
1	22k	100
2	22k	47
3	22k	33
4	22k	15

(RFB = 22k for all segments)

The basic filter networks are the same as for the Geigen Diapason (Figure 25a) but of course the component values are different:

Segment	R1	R2	R3	R4	C1	C2	C3
1	3k9	16k	75k	390k	220n	22n	4n7
2	1k6	9k1	68k	680k	220n	22n	2n2
3	2k2	11k	-	750k	100n	10n	-
4	2k2	10k	-	1M0	47n	3n3	-

(Level preset 100k)

Block Flute 2'

This stop is mainly intended to be used with the Stopped Diapason and/or Nason Flute though it has other uses with the swell flue work. It uses a pure square wave at 2 foot pitch:

Segment	SR1 (2')	C (nF)
1	22k	47
2	22k	33
3	22k	15
4	22k	6n8

(RFB = 22k for all segments)

The filter network circuit is at Figure 25b with components as below:

Segment	Cin	R1	R2	R3	R4	C1	C2	C3
1	100n	1k6	10k	68k	680k	220n	22n	2n2
2	68n	2k2	11k	-	1M0	100n	10n	-
3	33n	2k2	-	-	1M5	47n	-	-
4	15n	2k4	-	-	1M5	10n	-	-

(Level preset 100k)

Double Trumpet 16'

The waveforms for this stop have relative harmonic levels similar to a sawtooth:

Segment	SR1 (16')	SR2 (8')	SR3 (4')	C (nF)
1	22k	47k	82k	1000 (1uF)
2	22k	47k	82k	330
3	22k	47k	82k	220
4	22k	47k	82k	100

(RFB = 22k for all segments)

The filter networks are very different from any which have gone before. They use a parallel resonant circuit with a synthetic inductor, the theory of which was explained in the articles in *Wireless World* and EOM already referred to. Passive RC sections are also used to shape the response differently on each side of the resonant frequency. The basic circuit is in Figure 26a. The op amp used for the synthetic inductor could be a 741 but the TL071 is to be preferred. As there are four such resonators in the complete filter (one per segment) the quad version of this amplifier could be used if preferred (TL074). The amplifiers all operate off +/- 15 volts.

The components are as below, where R6 is the regulating resistor for each segment feeding into a TL071 summing amplifier as before. Also note that some values of capacitance are shown as being made up from two units. These are connected in parallel, not series, so that the capacitance values will add.

Segment	R1	R2	R3	R4	R5	R6	C1	C2	C3	C4
1	1k2	5k6	30k	68k	390R	82k	220n	22n	330n	22n
2	1k5	6k8	30k	47k	470R	130k	100n	15n	150n + 22n	15n
3	1k2	7k5	30k	82k	200R	100k	68n	15n	150n	6n8
4	1k0	4k7	30k	47k	470R	180k	47n	10n	68n + 10n	6n8

There may be audible breaks perceived between the segments of this stop, which are of little consequence in practice. However if desired they can be better disguised by regulating the relative levels of the segments by changing the R6's.

(Level preset 1M0)

Cornopean 8'

The harmonic proportions of the waveforms are sawtooth-like:

Segment	SR1 (8')	SR2 (4')	SR3 (2')	C (nF)
1	22k	47k	82k	330
2	22k	47k	82k	220
3	22k	47k	82k	100
4	22k	47k	82k	47

(RFB = 22k for all segments)

The filter networks are in Figure 26b, similar to the Double Trumpet but with an additional passive section:

Segment	R1	R2	R3	R4	R5	R6	R7	C1	C2	C3	C4	C5
1	4k7	18k	82k	82k	820R	56k	150k	22n	4n7	47n	10n	2n2
2	2k7	16k	75k	82k	820R	56k	220k	22n	4n7	33n	5n6	1n5
3	3k3	16k	75k	62k	1k1	62k	330k	10n	2n7	22n	4n7	1n0
4	3k3	15k	68k	91k	910R	62k	330k	10n	2n2	22n	2n2	680p

(Level preset 1M0)

Clarinet 8'

This stop uses the same waveform as the Stopped Diapason with mainly odd harmonics, hence the staircasing resistors etc. are the same. For convenience the component data are repeated below:

Segment	SR1 (8')	SR2 (4')	C (nF)
1	22k	180k	220
2	22k	180k	150
3	22k	180k	68
4	22k	180k	33

(RFB = 22k for all segments)

The filter networks are the same as for the Cornopean (Figure 26b) but of course with different component values. Component values are as below:

Segment	R1	R2	R3	R4	R5	R6	R7	C1	C2	C3	C4	C5
1	1k3	8k2	43k	82k	1k0	180k	180k	68n	10n	68n	3n3	470p
2	910R	6k2	36k	82k	1k0	130k	270k	68n	10n	47n	2n2	470p
3	910R	4k3	27k	82k	1k0	43k	680k	47n	10n	33n	1n5	1n0
4	910R	4k7	24k	82k	1k0	68k	1M0	33n	6n8	22n	1n0	470p

(Level preset 1M0)

Clarion 4'

The Clarion uses waveforms with similar harmonic proportions to a sawtooth:

Segment	SR1 (4')	SR2 (2')	C (nF)
1	22k	47k	220
2	22k	47k	100
3	22k	47k	47
4	22k	47k	22

(RFB = 22k for all segments)

The filter networks are similar to those of the Double Trumpet (Figure 26a):

Segment	R1	R2	R3	R4	R5	R6	C1	C2	C3	C4
1	1k2	11k	62k	47k	470R	82k	68n	10n	82n	10n
2	1k2	11k	62k	47k	470R	150k	47n	4n7	47n	6n8
3	1k5	10k	62k	51k	510R	180k	22n	3n3	22n + 6n8	3n3
4	2k2	10k	68k	51k	510R	160k	15n	2n2	22n	2n2

(Level preset 1M0)

Great Tone Filter Networks

Because the great keying system gives pitches up to 1 foot it is possible to derive a properly voiced and scaled Fifteenth, unlike on the swell, so the diapason chorus is thereby made more complete. Also the availability of this pitch gives slightly more high frequency edge to some combinations because it is used for the Principal as well.

Open Diapason 8'

This stop has an enhanced level of second harmonic like the Geigen Diapason, but it is derived and proportioned in a completely different manner. The enhancement occurs in the filter networks which use resonant filters, rather than in the driving waveforms as for the Geigen. This difference gives a distinctive flavour to this diapason, which is rounded with a "metallic" ring to the tone.

The waveforms have harmonic proportions similar to a sawtooth:

Segment	**SR1 (8')**	**SR2 (4')**	**SR3 (2')**	**C (nF)**
1	22k	47k	82k	330
2	22k	47k	82k	220
3	22k	47k	82k	100
4	22k	47k	82k	47

(RFB = 22k for all segments)

The filters are different to anything previously discussed. The circuit is in Figure 26c and it is a Sallen and Key second order Q-enhanced design. As with the other resonant circuits, the four op amps required should either be TL071's or a single TL074. Circuit values are as below where R3 is the regulating resistor:

Segment	**R1**	**R2**	**R3**	**C1**	**C2**
1	22k	6k8	180k	470n	10n
2	15k	5k6	270k	220n	10n
3	10k	4k7	390k	100n	10n
4	10k	5k1	1M0	47n	4n7

(Level preset 100k)

Claribel Flute 8'

This is a full toned flute with a quite different, more rounded, tone than the Stopped Diapason on the swell. It is useful as a solo stop, either on its own or accompanied from the swell. For this reason flutes of this type are less useful than the Stopped Diapason in combination with other stops, although it can support the Principal and Fifteenth as an alternative to the Open Diapason.

This stop uses the same waveforms as the Stopped Diapason, containing mainly odd harmonics as below:

Segment	**SR1 (8')**	**SR2 (4')**	**C (nF)**
1	22k	180k	220
2	22k	180k	150
3	22k	180k	68
4	22k	180k	33

(RFB = 22k for all segments)

The filter networks are as in Figure 25a :

Segment	**R1**	**R2**	**R3**	**R4**	**C1**	**C2**	**C3**
1	2k4	12k	56k	300k	470n	100n	22n
2	3k6	18k	91k	330k	220n	47n	10n
3	4k3	43k	220k	680k	100n	10n	2n2
4	5k1	51k	240k	1M5	47n	4n7	1n0

(Level preset 100k)

Dulciana 8'

This stop is a quiet diapason but it is different in tone from both the swell Geigen and the great Open, reflecting again the care taken with the design of the diapasons on this organ.

The waveforms have similar harmonic proportions to a sawtooth:

Segment	SR1 (8')	SR2 (4')	SR3 (2')	C (nF)
1	22k	47k	82k	330
2	22k	47k	82k	220
3	22k	47k	82k	100
4	22k	47k	82k	47

(RFB = 22k)

The filter networks are as in Figure 25a:

Segment	R1	R2	R3	R4	C1	C2	C3
1	1k5	5k1	22k	120k	330n	100n	22n
2	1k2	5k6	27k	220k	220n	47n	10n
3	2k2	18k	75k	470k	68n	10n	2n2
4	8k2	36k	75k	1M0	10n	2n2	470p

(Level preset 100k)

Principal 4'

As with all members of the diapason family on both manuals, care has been taken to ensure there is no "sameness" in tone either for the individual stops or when they are used in chorus. Thus considerable efforts were made when designing both the waveforms and the filters to get as close as possible to the sounds of real pipes of the highest quality. This stop has quite a different character from the Open Diapason and it is also distinctly different from the swell Geigens, yet its blending properties are not impaired.

The waveforms are derived as follows:

Segment	**SR1 (4')**	**SR2 (2')**	**SR3 (1')**	**C (nF)**
1	22k	47k	82k	220
2	22k	47k	82k	100
3	22k	47k	82k	47
4	22k	47k	82k	22

(RFB = 22k for all segments)

The filter networks are in Figure 25a, but not all of the components are used in all the segments. The lowest two use a 3rd order filter whereas the upper two are 2nd order. This is different from the Geigen Principal on the swell which uses 2nd order filters throughout:

Segment	**R1**	**R2**	**R3**	**R4**	**C1**	**C2**	**C3**
1	680R	3k3	15k	180k	330n	68n	15n
2	750R	3k6	22k	270k	220n	47n	4n7
3	1k0	4k7	-	360k	100n	22n	-
4	620R	3k0	-	430k	100n	22n	-

(Level preset 100k)

Fifteenth 2'

Again this is an individually tailored diapason stop to ensure optimum blend when used in chorus.

The waveform components are:

Segment	SR1 (2')	SR2 (1')	C (nF)
1	22k	47k	100
2	22k	47k	47
3	22k	47k	22
4	22k	47k	10

(RFB = 22k)

The filter networks are as in Figure 25c:

Segment	R1	R2	R3	C1	C2
1	330R	1k6	470k	330n	68n
2	390R	1k8	470K	220n	47n
3	470R	2k4	510k	100n	22n
4	300R	1k5	300k	100n	22n

(Level preset 100k)

Trumpet 8'

This stop is really a solo reed, designated Fanfare Trumpet in some organs I have designed, but if its volume is kept to a reasonable level it will also blend well with the full flue work of the great organ. In a small tonal scheme such as this it is more useful to have such a dual purpose stop on the great than a smoother or blander specimen.

The waveforms have the same harmonic proportions as a sawtooth:

Segment	SR1 (8')	SR2 (4')	SR3 (2')	C (nF)
1	22k	47k	82k	330
2	22k	47k	82k	220
3	22k	47k	82k	100
4	22k	47k	82k	47

(RFB = 22k)

The tone networks (Figure 26b) have been used for some of the swell reeds but the component values are different:

Segment	R1	R2	R3	R4	R5	R6	R7	C1	C2	C3	C4	C5
1	1k6	7k5	30k	82k	1k0	75k	68k	47n	10n	47n	10n	1n0
2	1k2	5k6	30k	82k	1k0	56k	100k	47n	10n	33n	6n8	1n0
3	820R	3k9	30k	82k	1k0	39k	240k	47n	10n	22n	4n7	1n0
4	1k2	5k6	30k	82k	1k0	56k	330k	22n	4n7	15n	3n3	470p

(Level preset 1M0)

Pedal Tone Filter Networks

In some ways pedal stops are the most difficult to design – preventing hootiness in the flue stops, and giving sufficient character to the 16 foot reeds so they penetrate and complement full organ without drowning it are but two of the problems. They also illustrate how skilful pipe design and voicing has to be in a first rate pipe organ. Many lessons have been learnt by carefully analysing the tonal structure of complete pedal departments of top quality pipe organs, rather than merely analysing the odd note here and there which is all that is usually done. These lessons have been implemented in the electronic designs which follow.

Because of the reduced compass the filters for the pedal stops only have two segments instead of four, with one exception described later.

Major Bass 16'

This stop is the loudest flue stop on the pedal. It uses waveforms with similar harmonic proportions to a sawtooth:

Segment	SR1 (16')	SR2 (8')	SR3 (4')	C (nF)
1	22k	47k	82k	2200 (2u2)
2	22k	47k	82k	680

(RFB = 22k)

The filter networks use the circuit of Figure 25a with component values as below:

Segment	R1	R2	R3	R4	C1	C2	C3
1	5k1	24k	100k	220k	680n	150n	10n
2	5k1	24k	100k	910k	330n	68n	4n7

(Level preset 470k)

Sub Bass 16'

This stop is the only one in the organ which uses only one filter segment, as it was impossible to improve on the effect by using more. However it is still necessary to use two sets of staircasing resistors for compatibility with the architecture of the system, but all the resistors (6 in all) can be fed into a single staircasing amplifier.

The driving waveform has similar harmonic proportions to a sawtooth:

Segment	**SR1 (16')**	**SR2 (8')**	**SR3 (4')**	**C (nF)**
1 & 2	22k	47k	82k	4700 (4u7)

(RFB = 22k)

The single filter network is in Figure 25a with the following component values:

Segment	**R1**	**R2**	**R3**	**R4**	**C1**	**C2**	**C3**
1 & 2	5k1	16k	68k	330k	1u0	220n	47n

(Level preset 1M0 shunted by 220pF)

Layout for this filter deserves some mention. Because of the extremely low harmonic development of the tones and their low frequencies it is important to keep the square wave inputs away from the input to the output amplifier, otherwise audible high frequency leakage will occur. The capacitor across the level preset helps to reduce any leakage signals which do break through.

Octave 8'

It was difficult to design the circuits for this stop in a way which prevented an unpleasant hootiness occurring (which sometimes also occurs in badly voiced pipes). However success was achieved by careful design of the tone networks for the two segments.

The waveforms have similar harmonic proportions to a sawtooth:

Segment	SR1 (8')	SR2 (4')	SR3 (2')	C (nF)
1	22k	47k	82k	470
2	22k	47k	82k	330

(RFB = 22k)

The filter networks are in Figure 27 with components as below:

Segment	R1	R2	R3	R4	R5	C1	C2	C3	C4
1	3k0	15k	68k	360k	820k	220n	47n	10n	4n7
2	2k2	15k	75k	-	3M0	100n	22n	4n7	-

(Level preset 220k)

Bass Flute 8'

A unique waveform was necessary for this stop with the following components:

Segment	SR1 (8')	SR2 (4')	SR3 (2')	C (nF)
1	22k	270k	560k	470
2	22k	270k	560k	330

(RFB = 22k)

The filter networks are in Figure 25a:

Segment	R1	R2	R3	R4	C1	C2	C3
1	3k6	18k	82k	390k	470n	100n	10n
2	1k1	9k1	68k	1M5	470n	100n	4n7

(Level preset 470k shunted by 470pF)

The same remarks about layout and sensitivity to breakthrough from the square wave inputs apply here as in the case of the Sub Bass.

Trombone 16'

In order to get the necessary effect of a telling reed without it being too powerful and thereby swamping everything else (a common problem even in pipe organs), it was necessary to tune the resonant circuits for this stop to harmonics of around 10 times the fundamental. This means that there is some attenuation of the fundamental, which can only be restored by adopting a filter configuration thought to be too complicated here. In practice the fundamental can be restored by using one or both of the 16 foot flue stops with the Trombone.

The waveform has the same harmonic proportions as a sawtooth:

Segment	**SR1 (16')**	**SR2 (8')**	**SR3 (4')**	**C (nF)**
1	22k	47k	82k	2200 (2u2)
2	22k	47k	82k	680

(RFB = 22k)

The filter networks are in Figure 26a with components as below:

Segment	**R1**	**R2**	**R3**	**R4**	**R5**	**R6**	**C1**	**C2**	**C3**	**C4**
1	1k3	6k2	68k	75k	430R	75k	220n	47n	100n + 68n	10n + 8n2
2	1k5	6k8	68k	68k	390R	120k	100n	22n	100n + 22n	10n + 2n2

(Level preset 1M0 variable in series with 680k)

Schalmei 4'

This stop is useful when playing Chorale Preludes. The waveforms have similar harmonic proportions to a sawtooth:

Segment	SR1 (4')	SR2 (2')	C (nF)
1	22k	47k	220
2	22k	47k	100

(RFB = 22k)

The filter networks are in Figure 26b:

Segment	R1	R2	R3	R4	R5	R6	R7	C1	C2	C3	C4	C5
1	470R	4k7	27k	82k	820R	47k	560k	220n	22n	47n + 33n	4n7	2n2
2	1k6	7k5	33k	82k	820R	68k	680k	47n	10n	47n + 10n	3n3	1n0

(Level preset 1M0)

CHAPTER 6

Output Channels, Amplifiers, Loudspeakers, Phase Shifters and Swell Pedals

Output Channels, Amplifiers and Loudspeakers

In the block diagrams for the swell, great and pedal departments (Figures 1, 18 and 21 respectively) each department was shown as having two output channels. Possible arrangements will now be discussed in more detail.

A recommended scheme is that in which the great and swell each have two independent output channels, making four in all, with a fifth for the 16 foot stops on the pedal. The 8 and 4 foot stops on the pedal would use one or other of the manual channels. (A "channel" means an independent power amplifier and loudspeaker system). The main advantage of this scheme is the opportunities it provides for enabling the mixing of different combinations of stops to occur acoustically in the listening room rather than electronically. The bonuses of such a scheme can hardly be over-estimated, because it gives much more liveliness to the tones of stops which are actually derived from a common generator system. It is essential, not just desirable, for acoustic mixing to occur for stops of the same family at different footages. Thus the two swell geigens must come from separate loudspeakers, as must the great open and principal. Similar remarks apply to the reeds.

The reason for this is that artificial reinforcement or cancellation of certain harmonics would otherwise occur if the mixing was done electronically prior to feeding the sounds to a single loudspeaker, owing to the phase shifts imposed by the tone filters. The practical consequence would be that when stops were combined the sound would not be what one would expect – a composite and artificial sound would result in which the careful voicing and regulation of the individual stops had been destroyed. The individuality of each stop as an acoustic entity becomes submerged by electronic mixing. When acoustic mixing occurs the worst effects of this phenomenon are avoided because the multiple reflections in the room break up the rigid reinforcement/cancellation scenario, and small movements of the player plus those of others in the room contribute to this desirable outcome.

The reason for requiring a separate channel for the pedal basses is rather different, and it is that these would otherwise result in severe intermodulation if fed into amplifiers and loudspeakers which also handle the outputs of higher pitch stops.

A five-loudspeaker system is not unduly elaborate or extravagant for an organ of this quality although its cost cannot be said to be trivial. The manual loudspeakers need to go down to about 60 Hz for the great, and desirably lower for the swell on account of its 16 foot reed, and good quality hi-fi units would be appropriate here. The loudspeaker for the pedal basses needs to have a really excellent bass response with a higher power handling capability, though this is not so easily provided. A single 15 inch or dual 12 inch loudspeaker units are recommended in a substantial bass reflex enclosure. A very successful and much cheaper alternative is to mount these units in the ceiling to capitalise on the large baffle thereby provided, which will often provide a response flat to below 30 Hz. (see my article in EOM 118, November 1985, p.26, reproduced here in Appendix 1).

It is convenient to use ordinary hi-fi amplifiers for the manual speakers as the tone controls on such units are useful for adjusting the overall effect of the organ in a particular room. It is often necessary to use some treble cut above 10 kHz or so to prevent the instrument sounding unduly spiky in a small room. In the same way bass lift below about 100 Hz can be useful for stops such as 16 foot reeds. The pedal channel on the other hand does not need tone controls, so it can be a simple integrated power amplifier without pre amplifier. Each amplifier should be of at least 60 watts for domestic use. Only a fraction of this would be used but this is normal practice if distortion is to be kept to a minimum, an important factor to prevent tiring the ear with the sustained tones of the organ.

A suggested scheme for distributing the stops to the various channels is given below. Virtual earth mixers using TL071's or TL074's should be used to combine the outputs of the respective tone filters into the associated channels.

Stop	**Sw chan 1**	**Sw chan 2**	**Gt chan1**	**Gt chan 2**	**Ped chan 1**
Geigen Diapason	*				
Stopped Diapason		*			
Geigen Principal		*			
Nason Flute	*				
Block Flute		*			
Double Trumpet	*				
Cornopean		*			
Clarinet	*				
Clarion	*				
Open Diapason			*		
Claribel Flute				*	
Dulciana			*		
Principal				*	
Fifteenth			*		
Trumpet				*	
Major Bass					*
Sub Bass					*
Octave				*	
Bass Flute			*		
Trombone					*
Schalmei			*		

If five channels are out of the question then a minimum of three should be used. Two would be implemented using hi-fi amplifiers with tone controls and hi-fi type speakers, and these would provide for the manual departments. The third would be for the pedal basses, with an extended bass response as outlined previously. A suggested stop distribution scheme in this case would be:

Stop	Manual chan 1	Manual chan 2	Ped chan 1
Geigen Diapason	*		
Stopped Diapason		*	
Geigen Principal		*	
Nason Flute	*		
Block Flute		*	
Double Trumpet	*		
Cornopean		*	
Clarinet	*		
Clarion	*		
Open Diapason		*	
Claribel Flute	*		
Dulciana		*	
Principal	*		
Fifteenth		*	
Trumpet	*		
Major Bass			*
Sub Bass			*
Octave	*		
Bass Flute		*	
Trombone			*
Schalmei		*	

Phase Shifters

An optional feature already referred to is the insertion of phase shifters in some of the outputs of the tone filters prior to combining them into the output channels. The signal phases would be slowly varied to impose a very small amount of frequency modulation on the signals. In effect this cyclically detunes them to disguise the fact that the stops for each department arise from a common generator system. Such an effect is often termed chorus, and together with the spatial separation of the stops obtained by using multiple loudspeakers, a very high degree of subjective independence for each stop can be achieved.

The use of phase shifters does not remove the need for multiple loudspeakers and for distributing the stops to different channels as already described. If this is not done the operation of the phase shifters becomes painfully apparent as the various stops drift in and out of phase lock. A most unpleasant “flanging” effect is heard as various harmonics become successively reinforced or cancelled, and the result can often be worse than if no phase shifters were used at all.

Phase shift should be applied to some of the stops that are often combined. Thus a phase shifter could be inserted in the output of the Geigen Diapason so that it will be unlocked when used either with the Geigen Principal or the Stopped Diapason. To illustrate the point further the table below is a reproduction of the previous one, where the symbol P means that a phase shifter is used on that stop:

Stop	Sw chan 1	Sw chan 2	Gt chan1	Gt chan 2	Ped chan 1
Geigen Diapason	P				
Stopped Diapason		*			
Geigen Principal		*			
Nason Flute	*				
Block Flute		*			
Double Trumpet	*				
Cornopean		P			
Clarinet	*				
Clarion	*				
Open Diapason			*		
Claribel Flute				*	
Dulciana			*		
Principal				P	
Fifteenth			*		
Trumpet				*	
Major Bass					*
Sub Bass					*
Octave				*	
Bass Flute			*		
Trombone					*
Schalmei			*		

In this example 3 phase shifters are needed, and they can be moved around the system until the best subjective effect is obtained. Providing too many will result in the flanging effect becoming too obvious, as there are so few independent loudspeaker channels available hence few opportunities for acoustic as opposed to electronic mixing. Flanging is most objectionable when electronic mixing is used but it is usually undetectable with acoustic mixing. The use of a few phase shifters will result in a just-perceptible sense of "movement" in the sound of the organ to compensate for the use of a common generator system for each department.

I described a suitable circuit using a field effect transistor for each phase shifter in EOM 178 (July 2001), reproduced here at Appendix 2. Some years ago I also designed a more sophisticated system using bucket brigade delay lines which has been used by various members of the society. Unfortunately the TDA 1022 delay line used in this design is long obsolete although alternative components are still widely available. One example is the MN3207 (Maplin type UR67X) which has a better performance in several respects than the TDA 1022. Design details of the original system appear here at Appendix 3, but it must be emphasised that some redesign will be necessary to accommodate today's bucket brigade devices.

Swell Pedals

A dual channel swell control is required for the swell department, and this should have certain desirable characteristics if the quality of the instrument is not to be compromised. Firstly there should be a tone control effect so that the higher frequencies are attenuated more than lower ones, as with a real swell box. Secondly the control should incorporate some form of time constant so that no matter how fast the pedal is operated, the sound grows or decays in a realistic manner. This is necessary because even with a mechanically operated swell box in a pipe organ, there is a limit set by inertia on how fast one can operate the pedal. With a pneumatic or electric connection there is always some time delay involved. Thirdly the variation of volume with pedal movement should be most critical near to the fully closed position. Fourthly the circuit should be able to control as many channels as necessary – at least two in the present instrument.

A circuit which incorporates all these features was described in my articles in EOM 124 and 125 (February and April 1987), reproduced here at Appendix 4. Unfortunately the MC1495L analogue multiplier used in this design is now hard to obtain. However other analogue multipliers are available. Also a simplified form of the circuits using the readily available CA3080 transconductance operational amplifier was suggested by David Ryder in EOM 132 (February 1989), although I have not tried this circuit myself..

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CHAPTER 7

Key Contacts and Couplers

So far it has been assumed tacitly that each octave of keying circuits was connected directly to the respective key contacts using 16 way ribbon cables. This would be fine for a single manual organ, but where couplers are required some further complication is necessary.

Let us first get multiplex keying out of the way. For a single keyboard, multiplex keying is merely a complicated method of connecting 61 contacts to 61 keying circuits and it has nothing to recommend it. Only when there are couplers does multiplexing result in economies, mainly in construction time. A more detailed analysis suggests that a break-even point occurs regarding component count and assembly time when there are about three couplers, as in this instrument. Consequently multiplex keying will not be described here, and another reason is that many articles have appeared on the subject in EOM. An incomplete list includes articles by Ryder (nos. 98 – 104), McGrechan (no. 123), Hawkins (nos. 128, 132, 134, 146 & 163) and Harris (no. 130). However if you want to use multiplex keying then the 16 way ribbon cables from the keyers will simply plug in to your system, which must be able to apply 15 volts to the keying circuits to switch a note on.

Key contact materials are not critical because of the use of 15 volt levels for keying. Thus silver wires are perfectly satisfactory and they will continue to operate even when tarnished and thoroughly black. This would not have been the case had lower voltages such as 5 volts been used. Gold clad wires will of course also function well. Baser metals such as phosphor bronze might also give satisfactory service though it has not been tried with this particular keyer.

Couplers

There are two basic types of coupler circuit depending on whether you want to use only a single key contact per note regardless of how many couplers are used, or whether you are prepared to use more contacts. A circuit for three notes of the single contact per note scheme is in Figure 28. The switches enclosed in rectangles are the coupler switches, and 4066 analogue switch IC's are recommended as they are cheap and compact with only 15 packages being required for a swell to great coupler (16 if the top note is to be included on a 61 note keyboard). The coupler is switched on by setting all the enable (E) pins of the 4066's high. When the coupler is off the 100k resistor pulls the enable inputs to ground. This is shown for the swell to great, and the other couplers have identical switching arrangements. The Y and Z terminals of the switch are interchangeable so it does not matter which way round it is connected.

The other switch symbols not enclosed in rectangles denote the key contacts. Note that the couplers "couple through" in this scheme i.e. if the swell to great is on, then the great *and* the swell will sound when only the great to pedal is drawn. This seems the more logical plan as it more accurately simulates what would happen in a pipe organ with mechanical coupling. The diodes are essential to prevent reverse coupling i.e. the swell to great must not also act as a great to swell.

A diagram for the multiple contact per note scheme is in Figure 29. This has the advantage that the 61 way electronic relay required for the swell to great (constructed from 4066's in the former scheme) is not required. Similarly, the 30 or 32 way relays for the pedal couplers are not needed either. Nor does the swell "couple through" in the manner described earlier. However it is necessary to provide 2 contacts per note for the great, and three for the pedals. Thus the key wiring and contact arrangements get correspondingly more complicated, although careful design and layout using ribbon cable and IDC connectors will reduce the tedium and labour involved very considerably.

Each key contact connects to a common line which is permanently at 15 volts for the contacts of the corresponding department. The common lines for the coupled departments only receive 15 volts when the associated coupler is drawn. The use of commercial key contact blocks is recommended here, such as those made by Kimber-Allen. The swell would need single pole contacts (K-A ref GJ) and the great 2 pole (ref GB). The pedals could use 3 pole contacts of similar design (ref GC) provided care is taken to ensure that the greater movement of the pedal keys does not unduly strain the gold wires. This could be ensured by using proper pedal jacks to actuate the contact assemblies. The "wiper" for all contacts of the K-A type is better regarded as an actuator in this application, as it must be insulated.

The above circuits are but two of several ways of coupling the keyboards. Even as late as the 1980's some of the finest commercial analogue organs used entirely standard pipe organ console arrangements, with electro-mechanical coupler relays. Other more sophisticated electronic arrangements can be devised as alternatives to those described. It has merely been the intention of this chapter to indicate some ways in which coupling can be achieved.

CHAPTER 8

Free Phase Organs

In previous chapters circuit details were given of a divider organ, concentrating on novel tone generation and keying circuits using standard integrated circuits. This chapter begins a discussion of free phase (FP) tone generation.

The most important requirement for a FP oscillator is that it must be highly stable; the need for frequent retuning when there are hundreds of oscillators would be intolerable. For musical purposes a long term frequency stability of at least 0.05% is necessary (500 parts per million) over periods when the ambient temperature can fluctuate easily by 20 ^{o}C even in domestic rooms. Thus about 25 ppm per ^{o}C is needed. So-called high stability passive components (resistors, capacitors or inductors) seldom have a better long term temperature coefficient than 50ppm per degree, so with only one of them in a circuit this figure would barely be adequate for long term stability. When there are several components as in a practical oscillator all drifting independently, the magnitude of the problem can be appreciated. There are also ageing problems with components which manufacturers seldom quantify. Without going into a lot of detail all this means that the oscillators must be designed to produce sine waves. Good frequency stability implies an oscillator which will be reluctant to oscillate at any but a specific frequency, and this means a sine wave has to be produced which only has a single Fourier component. Designing an oscillator to produce complex waveforms directly, such as sawtooths or square waves which contain many Fourier components, implies some form of nonlinear semiconductor circuitry which itself is invariably temperature dependent or which relies on power supplies which can seldom attain the degree of stability required. (Having said that, a number of members have used 555 type timer chips as free phase oscillators which produce complex waves directly, though I have little personal experience of this type of system).

Unfortunately sine waves are of little value for tone forming, except perhaps for rather uninteresting flute-like tones. They can however be shaped in various ways, so attention is focused here on converting the sine wave output of the oscillators into sawtooth and square waves which then can be used with the tone filter networks already described. In fact no use at all is made of the sine waves in what follows. However another way of shaping sines is to clip them, and David Ryder pursued this in some detail (see EOM 100, April 1982 *et seq*). This approach led to some attractive diapason-like voices together with a range of “fat” and “thin” reeds. Yet another way to use sine waves is to inject them selectively into tone filters where an augmented fundamental is required, larger than the use of a sawtooth or square wave would permit.

There are several ways to design a FP organ. The most basic though not necessarily the simplest is to have just one rank of FP oscillators which feed the entire instrument, or at least a single department. If the pitch range of the organ extends from 16 feet to 2 feet then 8 octaves or 96 oscillators would be needed. However a complicated switching system is needed to distribute the oscillator outputs to the various footages, and this requires wiring and analogue switching comparable to that of a divider organ.

In fact the two schemes are similar except that in the divider system the single rank of generators is running all the time. In the FP case the keying can be more difficult than for dividers if the natural attack and decay of the oscillators is not to be degraded by the key switches.

Another scheme is to have an independent rank of oscillators for each footage. Although many more oscillators are required it is attractive because the keying becomes much simpler. Therefore the active devices in the keying system of the first scheme can be used to generate additional independent frequencies rather than merely routeing them around the system, which is a much better use of resources.

Yet another scheme is to have an independent set of oscillators per stop. This involves as many oscillators as pipes in a pipe organ with the same stop list. There can be no doubting the sheer musicality of this approach, but it is seldom a practical proposition.

The system chosen here is the second one in which there is an independent rank of FP oscillators per footage per department. Perhaps surprisingly, a compelling reason for considering this scheme lies in its economy regarding the number of IC packages required. Considering first the divider system already discussed, the swell requires about 150 IC packages plus 61 transistors per manual department (i.e. including the two TOGs and the keying/divider circuits). A FP system with about 240 oscillators to produce the same four footages would need only about 60 quad operational amplifiers. However depending on the type of oscillator, each might need a keying transistor. If quad analogue switches were used for keying instead, then another 60 packages would result, giving a total package count of about 120. This is still less than for the divider system and no discrete transistors would be needed, yet all the advantages of free phase would be obtained – there would be absolutely no phase locking between notes of a given footage, nor between notes at different footages. Each note at each pitch would automatically have an attack and decay related to its frequency, determined by the frequency and Q of the associated oscillator. The instrument could be tuned to any desired temperament, and because each oscillator could provide a sawtooth wave some advantageous tonal changes could be made to the stop list of the organ (e.g. the swell could have a Fifteenth instead of just a 2 foot flute). There would be no beehive.

For the great department only three ranks would be needed because there are only three footages, therefore even fewer packages would be required compared to the divider system. At first sight this may seem obscure, as the divider system needed four footages. But this is because the dividers only generate square waves, and for the 2 foot stop it was necessary to add (via staircasing) the even harmonics from a 1 foot rank so that a proper Fifteenth could be voiced. With the FP system which would provide sawtooth waves with a full retinue of harmonics, this would be unnecessary. Similar remarks apply to the pedal department.

What are the downsides of a FP system? Clearly it is not straightforward to build and set up a system with so many independent oscillators (though certainly no more difficult than tuning a pipe organ!). Also there would be less flexibility in tone forming if only square and sawtooth waves were available - one advantage of the divider system is the ability to build a variety of waveforms merely by altering the staircasing resistors which feed the tone networks. Each FP oscillator needs typically about a dozen passive components (this includes the sawtooth and square wave shapers) in addition to the active devices. This is a disadvantage in itself, and it also contributes to some doubt about tuning stability. It would have to be accepted that a FP organ would probably need regular if not frequent tuning in the same way as a pipe organ. Nevertheless it is possible to make too much of this shortcoming, as very large FP organs have been made in the past in which tuning was not the main problem. Some of them were unreliable in other ways simply because of the huge number of components.

On balance there are strong arguments for regarding a FP organ as a genuine musical instrument, and which does not necessarily need any more components than one which uses dividers, and it is on this basis that the following design notes are offered. The block diagram of the swell department for an organ built along these lines is in Figure 30, the great in Figure 31 and the pedal in Figure 32.

In this design, when a note is keyed all oscillators for that note in that department are activated regardless of whether the corresponding stops are drawn. Thus if a note on the swell was keyed the 16, 8, 4 and 2 foot oscillators would all be switched on, but the only sounds heard would be those for which the stops were drawn. This makes for an extremely simple keying system. Purists might argue that this could result in an increase in the background noise of the system, because an advantage of FP is that oscillators need not be switched on until they are actually needed. This is a valid argument in theory, but in practice an audible increase in background noise would only occur if the system was badly designed with the most appalling opportunities for breakthrough. It would not be out of the question to gate the oscillator keying with the stops actually drawn, although this would considerably complicate the keying arrangements.

Each oscillator has wave shaping circuitry to convert its sine wave output into sawtooth or square waves, or both. In Figure 30 it is seen that sawtooths only are used for the 16 foot rank, both sawtooths and squares for the 8 and 4 foot ranks, and squares only for the 2 foot rank. A different arrangement appears for the great in Figure 31. These requirements arise purely because of the tone forming required for the particular stop list used in this organ and it can of course be varied. The keyboard is segmented as before into four segments for tone forming purposes (two segments on the pedals). In each segment, all the outputs of the sawtooth wave shapers are mixed in a common amplifier and similarly for the square waves. These mixers feed a bus arrangement as indicated in the diagrams, which are drawn so that each bus (indicated by the bold lines) actually consists of the four outputs from the corresponding segments.

The tone forming filters pick off sawtooth or square waves from each segment as shown. Because there is no need for staircasing amplifiers each tone filter is simpler than in the case of the divider organ previously described, but the penalty paid is the lack of flexibility in waveform design which will somewhat degrade the fidelity of tone forming in this case for a few stops (though there is a way to solve this problem which is described later). The filters feed dual output channels, via phase shifters in some cases, in much the same manner as before. However in this FP scheme all stops arise from independent generators except for those of the same footage. Therefore the main function of the phase shifters in this case is to break up the phase locking which occurs between these stops only. Consequently there are some minor differences between the allocation of phase shifters and output channels for the divider and FP schemes.

A major difference between the swell department for the FP scheme compared with the divider scheme is that the tremulant has to be applied after the tones have been formed, rather than by modulating the frequencies of the oscillators. There are many ways to do this, and the choice depends partly on whether one wants amplitude or frequency modulation of the sounds. Most people agree that the latter is preferable and I described a simple circuit in EOM 178 (July 2001) using FET's, reproduced here at Appendix 2. An alternative approach using bucket brigade delay lines was described by David Ryder in EOM 87 (September 1979) *et seq*.

CHAPTER 9

Free Phase Oscillators

Even when the decision has been taken to only consider sine wave oscillators, there is still a large number of options. The obvious one is to use LC oscillators with ferrite cored coils which can also be used for tuning. However these are expensive and the labour of winding the coils, particularly if they are centre tapped, is considerable. Therefore in this design the use of coil-less free phase (CFP) oscillators will be described.

The CFP oscillator due to David Ryder is properly and elegantly designed and it has been used in large quantities by a number of constructors. It was described in EOM 97 (September 1981) and readers are referred to this for complete details which it would be improper to reproduce here. Only those details which are necessary for the successful integration of this oscillator with the architecture of this FP organ will be summarised. The circuit with minor elaborations to suit this organ is in Figure 33, and the main features of the oscillator itself are (with acknowledgments to Dr Ryder):

1. The two resistors R1 and R2 are equal and their value is calculated for a specific frequency by the formula:

$$R1 = R2 = R = \frac{\left\{\frac{159155}{f} - 2.39\right\}}{\sqrt{C1.C2}}$$

where f is frequency in Hz, C1 & C2 are in nF and R in kohm.

2. Q is determined by the ratio of the capacitors and it determines the attack and decay rate of the oscillator. The ratio shown in the diagram gives an excellent natural attack and decay across the entire compass to my ears, from 16 foot C1 to 2 foot C6, though some may find it gets too slow at the lower frequencies. The values of capacitance shown can be maintained until R starts to exceed about 1 Mohm, at which point they can be increased. However the ratio should remain the same, subject to the previous remark. These matters are discussed in detail in the article referred to.

3. Keying is done by opening the analogue switch, giving the natural envelope referred to above. A discrete transistor switch was suggested by Dr Ryder but the use of the 4066 may make for a more compact board layout as only one package suffices for the one quad op amp (a TL074) and hence four oscillators. Note that keying is done by making a contact to 0 volts rather than to 15 volts as in the divider system. As many oscillators can be controlled as necessary from the one key contact but each needs its own analogue switch section.

4. Diodes D1 and D2 are actually part of the oscillator and they stabilise its amplitude, and the square wave signal can be conveniently picked off from

this point. However the diodes must remain even if no square wave output is required from a given rank.

5. The sawtooth output is obtained using a differentiator followed by a diode. These components can be omitted when no sawtooth is required. It is essential to keep the waveshape (hence relative harmonic amplitudes) constant, therefore the values of R3 and C3 vary with frequency. Complete values are given in the tables which follow. More detail about this and other types of nonlinear wave shaping was given by Dr Ryder in EOM 106 (June 1983).

16’ 8ve 1	C3	R3	16’ 8ve 2	C3	R3	16’ 8ve 3	C3	R3	16’ 8ve 4	C3	R3	16’ 8ve 5	C3	R3
C	470	20	C	220	22	C	100	24	C	47	24	C	33	18
C#	470	20	C#	220	20	C#	100	22	C#	47	24	C#	22	24
D	330	24	D	220	20	D	100	22	D	47	22	D	22	24
D#	330	24	D#	220	18	D#	100	20	D#	47	22	D#	22	22
E	330	22	E	150	24	E	100	20	E	47	20	E	22	22
F	330	22	F	150	24	F	100	18	F	47	20	F	22	20
F#	330	20	F#	150	22	F#	68	24	F#	33	24	F#	22	20
G	330	20	G	150	22	G	68	24	G	33	24	G	22	18
G#	330	18	G#	150	20	G#	68	22	G#	33	22	G#	15	24
A	220	24	A	150	20	A	68	22	A	33	22	A	15	24
A#	220	24	A#	150	18	A#	68	20	A#	33	20	A#	15	22
B	220	22	B	100	24	B	68	18	B	33	20	B	15	22

8' 8ve 1	C3	R3	8' 8ve2	C3	R3	8' 8ve3	C3	R3	8' 8ve 4	C3	R3	8' 8ve 5	C3	R3
C	220	22	C	100	24	C	47	24	C	33	18	C	15	20
C#	220	20	C#	100	22	C#	47	24	C#	22	24	C#	15	18
D	220	20	D	100	22	D	47	22	D	22	24	D	15	18
D#	220	18	D#	100	20	D#	47	22	D#	22	22	D#	10	24
E	150	24	E	100	20	E	47	20	E	22	22	E	10	24
F	150	24	F	100	18	F	47	20	F	22	20	F	10	22
F#	150	22	F#	68	24	F#	33	24	F#	22	20	F#	10	22
G	150	22	G	68	24	G	33	24	G	22	18	G	10	20
G#	150	20	G#	68	22	G#	33	22	G#	15	24	G#	10	20
A	150	20	A	68	22	A	33	22	A	15	24	A	6.8	27
A#	150	18	A#	68	20	A#	33	20	A#	15	22	A#	6.8	27
B	100	24	B	68	18	B	33	20	B	15	22	B	6.8	24

<u>Tables showing values of sawtooth converter components for 16’ & 8’ CFP ranks</u>

(C3 in nF; R3 in kohm)

4' 8ve 1	C3	R3	4' 8ve2	C3	R3	4' 8ve3	C3	R3	4' 8ve 4	C3	R3	4' 8ve 5	C3	R3
C	100	24	C	47	24	C	33	18	C	15	20	C	6.8	22
C#	100	22	C#	47	24	C#	22	24	C#	15	18	C#	6.8	20
D	100	22	D	47	22	D	22	24	D	15	18	D	6.8	20
D#	100	20	D#	47	22	D#	22	22	D#	10	24	D#	6.8	18
E	100	20	E	47	20	E	22	22	E	10	24	E	4.7	24
F	100	18	F	47	20	F	22	20	F	10	22	F	4.7	24
F#	68	24	F#	33	24	F#	22	20	F#	10	22	F#	4.7	22
G	68	24	G	33	24	G	22	18	G	10	20	G	4.7	22
G#	68	22	G#	33	22	G#	15	24	G#	10	20	G#	4.7	20
A	68	22	A	33	22	A	15	24	A	6.8	24	A	4.7	20
A#	68	20	A#	33	20	A#	15	22	A#	6.8	24	A#	3.3	24
B	68	18	B	33	20	B	15	22	B	6.8	22	B	3.3	24

2' 8ve 1	C3	R3	2' 8ve2	C3	R3	2' 8ve3	C3	R3	2' 8ve 4	C3	R3	2' 8ve 5	C3	R3
C	47	24	C	33	18	C	15	20	C	6.8	22	C	3.3	22
C#	47	24	C#	22	24	C#	15	18	C#	6.8	20	C#	3.3	22
D	47	22	D	22	24	D	15	18	D	6.8	20	D	3.3	20
D#	47	22	D#	22	22	D#	10	24	D#	6.8	18	D#	3.3	20
E	47	20	E	22	22	E	10	24	E	4.7	24	E	3.3	18
F	47	20	F	22	20	F	10	22	F	4.7	24	F	2.2	24
F#	33	24	F#	22	20	F#	10	22	F#	4.7	22	F#	2.2	24
G	33	24	G	22	18	G	10	20	G	4.7	22	G	2.2	22
G#	33	22	G#	15	24	G#	10	20	G#	4.7	20	G#	2.2	22
A	33	22	A	15	24	A	6.8	24	A	4.7	20	A	2.2	20
A#	33	20	A#	15	22	A#	6.8	24	A#	3.3	24	A#	2.2	20
B	33	20	B	15	22	B	6.8	22	B	3.3	24	B	2.2	18

Tables showing values of sawtooth converter components for 4’ & 2’ CFP ranks

(C3 in nF; R3 in kohm)

Tuning

David Ryder's recommendation was to pre-tune the oscillators by measuring and sorting the capacitors (C1 and C2) and then making up resistors to suit from the E24 range. This is the cheapest and probably the most stable approach but any retuning that might be required, perhaps due to ageing of components, could be rather fiddly. It is also somewhat time consuming in the construction phase. Therefore some form of variable tuning might be preferred, the two obvious choices being a preset resistor in series with one of the fixed resistors, or a trimmer across one of the capacitors.

Although this would undoubtedly make the oscillators more convenient to construct and retune it introduces two disadvantages. Trimmers have a temperature coefficient up to 10 times worse than the best fixed capacitors, and those of preset resistors are typically 2 or 3 times worse than fixed resistors. The second disadvantage is cost – the good quality stable trimmers or presets that are needed will cost in the region of £1 each at 2001 prices, and this will dominate the cost of the oscillator components. But on the assumption that one is likely to want to pay more for the convenience of having a system that one could tune, then the choice is between variable capacitors or resistors. Although trimmers have a worse temperature coefficient, it acts in the reverse direction to that of the resistors in the circuit, and this may also apply to the fixed capacitors as well depending on their type. So it is possible the oscillator as a whole could be quite stable using a capacitive trimmer to tune it.

The tuning components could either be 22 pF miniature film trimmers across the 100 pF fixed capacitors (Maplin type WL70M) or multi-turn cermet presets in series with one of the resistors. The latter option is less convenient as the value of the preset needs to be related to that of the fixed resistor which it augments and this will depend on frequency, whereas the trimmer value can remain constant for much of the frequency range, i.e. until the value of the 100 pF capacitor needs to increase, at which point a larger trimmer would be needed. There is little to choose between the two approaches as regards cost. If the capacitive trimmer option is adopted then the oscillator frequency should be designed for a value of C1 = 111 pF (trimmer half open), and the two fixed resistors then made up from at least two E24 units in series to get as close as possible to the desired value.

Mixers

The waveforms from the oscillators are combined for each keyboard segment in virtual earth mixers as in Figure 34, giving outputs at about 2 volts peak-to-peak for each note keyed. It will be noted that the arrangement for sawtooths is slightly different to that for the square waves - the feedback resistors are different because of the different amplitudes of the two waveforms arising from the oscillators. Also the square waves are symmetrical with respect to ground, but the sawtooths are not. This means that there is a net DC potential associated with the sawtooth waves which could give rise to keying thump if not removed. This would be most noticeable in the higher frequency regions of stops with low harmonic development such as flutes. DC removal is done in the tone filters as for the divider system, and it will be described presently.

Note also the different arrangements for segment 4, which has to collect the outputs from octave 5 as well as octave 4. 13 additional resistors are needed in this case. The resistors for octave 5 are also less than those for the other octaves for the same reason as in the divider organ case: to provide some additional boost to signals in the top octave.

Because each note at each pitch in this organ has its own regulating resistor (i.e. the 220k or 150k resistor feeding the mixers) it is simple to regulate the odd note which may need it in a particular room. This might be necessary to compensate for room or loudspeaker characteristics.

Constructional Notes

It is suggested that a separate board is constructed for the oscillators and mixers for each octave of each footage. The twelve oscillators can be built round three TL074 packages, and the keying circuits will use a further three 4066's. Provision should be made for each of the resistors R1 and R2 to be made up from at least two units. One TL071 mixer will be required for the square wave outputs and another for the sawtooths, but recall that not all footages need both (refer to Figures 30, 31 & 32). Thus a maximum of 8 packages are needed per octave to produce 12 notes with two waveforms per note. For the top octave no mixers are needed as the waveform outputs are mixed on the octave 4 boards as described already. This will involve running flying leads from one board to the other, and they should be kept as short as possible in view of the relatively high impedance at this point in the circuit. There will be no danger of picking up beehive because no oscillators run until they are keyed, but other unwanted signals such as hum or even RF transmissions could conceivably be a problem if these leads are too long.

If the top note (C6) is required an additional TL071 and one section of a 4066 is needed in addition to the components for the other 12 notes – this illustrates again how awkward the top note can be, and you may wish to omit it.

The power lines coming in to each board should be decoupled with 47 ohm resistors and 100 uF capacitors, and the ground lines should be of relatively heavy gauge wire.

A photograph of a PCB containing one octave of CFP oscillators as used at one time by a commercial organ company is on the front cover of EOM 125 (April 1987).

Output Waveforms

The waveforms at the outputs of the mixers are not in fact perfect square or sawtooth waves and this deserves some comment. The "square" waves arise from passive diode limiters and they have rounded tops and bottoms as illustrated in Figure 35. The sawtooths not only have exponential rather than linear slopes, but they are only half-sawtooths occupying only half the inter-period interval.

Perhaps surprisingly these differences from normal waveforms hardly matter at all. The harmonic spectra of these waveforms were measured, and these showed that the "square" waves are equivalent to ordinary square waves up to about the 12th harmonic, after which the harmonic amplitudes fall off rather more rapidly by an additional 6 dB per octave. The "sawtooth" waves are identical to perfect sawtooths up to the 17th harmonic, after which there is again an additional fall off of 6dB per octave.

The tone forming implications will hardly be noticeable for those filters which use mainly low pass networks, and this means all of the flue stops. The same will be true for the reeds which use resonant sections with the possible exception of 8 and (particularly) 16 foot reeds lower in the compass, where they might possibly sound smoother with these waveforms compared to those used in the divider organ. However it is doubtful whether such differences could be detected other than by deliberate A-B comparison tests. Nevertheless some minor changes will be suggested to certain filters shortly.

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CHAPTER 10

Tone Forming for CFP Oscillators

The tone forming system described for the divider organ in chapter 5 can be used with few changes when using CFP oscillators. Referring to Figure 36, buffer amplifiers (previously they were staircasing amplifiers) at the input of each filter segment are still present, and they also retain the DC blocking capacitors to remove the DC component present in the sawtooth waves. The value of the capacitor should be calculated using the formula on page 25. The values of R1 and R2, on which the capacitor values depend, are given later for each stop. Normally each amplifier would pick off a single waveform from either the square or the sawtooth busses, but in a few cases it is advantageous to use both waveforms together and this is indicated in the diagram. This enables a controlled amount of even harmonics to be injected into tones which require mainly odd ones.

If this refinement is not required then certain tones will not be formed with the fidelity which the divider system offered. However the practical advantage is that the buffer amplifiers could then be dispensed with altogether, and the filter networks for each segment connected directly to the corresponding square or sawtooth bus. Unfortunately this might turn out to be a false economy as the input impedance of some networks is quite low, and several of them connected directly to a bus might load it too heavily. In these circumstances the mixer amplifier which feeds the bus would then introduce intermodulation distortion when more than one note was keyed. Also the blocking capacitor cannot be omitted when sawtooths are used. On balance, therefore, it is recommended that the buffer amplifiers are retained.

<u>Notes on Tone Forming</u>

1. The Geigen Diapason requires a waveform which, in most segments, has the 2nd harmonic amplitude equal to the fundamental. This was possible with the flexibility offered by the divider system with its staircasing options, but in the free phase case it is not. Therefore it is suggested a sawtooth wave is used for this stop, which will result in a somewhat less bright and interesting sound.

2. The following stops also require sawtooth waveforms, but in this case no tonal degradation should occur. Only one input resistor (R1) to each buffer amplifier in Figure 36 is needed, and this should have a value of 22k:

Geigen Diapason (see note above), *Geigen Principal, Double Trumpet, Cornopean, Clarion, Open Diapason, Dulciana, Principal, Fifteenth, Trumpet, Major Bass, Sub Bass, Octave, Trombone, Schalmei*

3. The following stops require square waveforms. Only one input resistor (R2) to each buffer amplifier in Figure 36 is needed, and this should have a value of 22k:

Nason Flute, Block Flute

4. The following stops could use square waves only as they require mainly odd harmonics. However better results will be obtained by using a special waveform

consisting of mainly odd harmonics with a lower level of evens. It is derived by using square and sawtooth waves simultaneously and setting R2 = 22k and R1 = 91k in Figure 36. This gives a reasonable approximation to the harmonic amplitudes obtained in the divider organ by staircasing:

Stopped Diapason, Clarinet, Claribel Flute

5. The Bass Flute also could use a square wave only, but it would benefit from another type of waveform in which there is a lower level of even harmonics than those present in the former stops. It is derived by using square and sawtooth waves simultaneously and setting R2 = 22k and R1 = 130k in Figure 36.

6. Because 2 foot sawtooths are available in the free phase system it is possible to add a diapason – quality stop to the swell at this pitch, rather than having a flute only. This would enhance the diapason chorus on this department considerably. Similar tone networks to those for the great Fifteenth should be used.

7. Because of the slightly reduced amplitudes of the upper harmonics in the sawtooth waves, minor changes might be tried to a few tone filters. C2 for the Trombone (Figure 26a) might be removed in both segments, which will produce a slightly brighter sound. Similarly, C1 might be removed from the lower segments of the Double Trumpet. These adjustments could also be tried for other reeds in the lower segments, but note that C3 and C4 should not be touched in any of the reed filters if the tones as designed are to be retained. These determine the characteristics of the resonant sections and are fairly critical.

CHAPTER 11

Free Phase Organ - Miscellaneous Topics

This chapter discusses various topics which were also covered for the divider organ, concentrating on areas where there are differences.

Output Channels, Loudspeakers, Phase Shifters and Swell Pedals

Everything mentioned in earlier chapters applies here, the only differences arising from the additional diversity or degrees of freedom in a free phase oscillator system compared to dividers. In each department, all stops except those of the same footage arise from independent oscillators which are not phase locked. It is still highly desirable to distribute the stops to separate output channels following the same rationale described in chapter 6. However in this case the only stops which need to be unlocked are those of the same footage, and phase shifters can be applied to a few of them. A suggested scheme is in the block diagrams (Figures 30, 31 and 32).

Key Contacts and Couplers

Again, the discussion in chapter 7 largely covers the requirements for free phase oscillators. Each key contact in this case is connected to the enable pins of as many 4066 switches as there are ranks of oscillators in a department. However an important point is that keying is done by grounding these pins, rather than by connecting the keying lines to 15 volts as with the divider organ. This means that all diodes in the coupler circuits given previously must be reversed. (NB: do not get confused between the 4066's which switch the oscillators, and those used in the couplers).

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DIAGRAMS

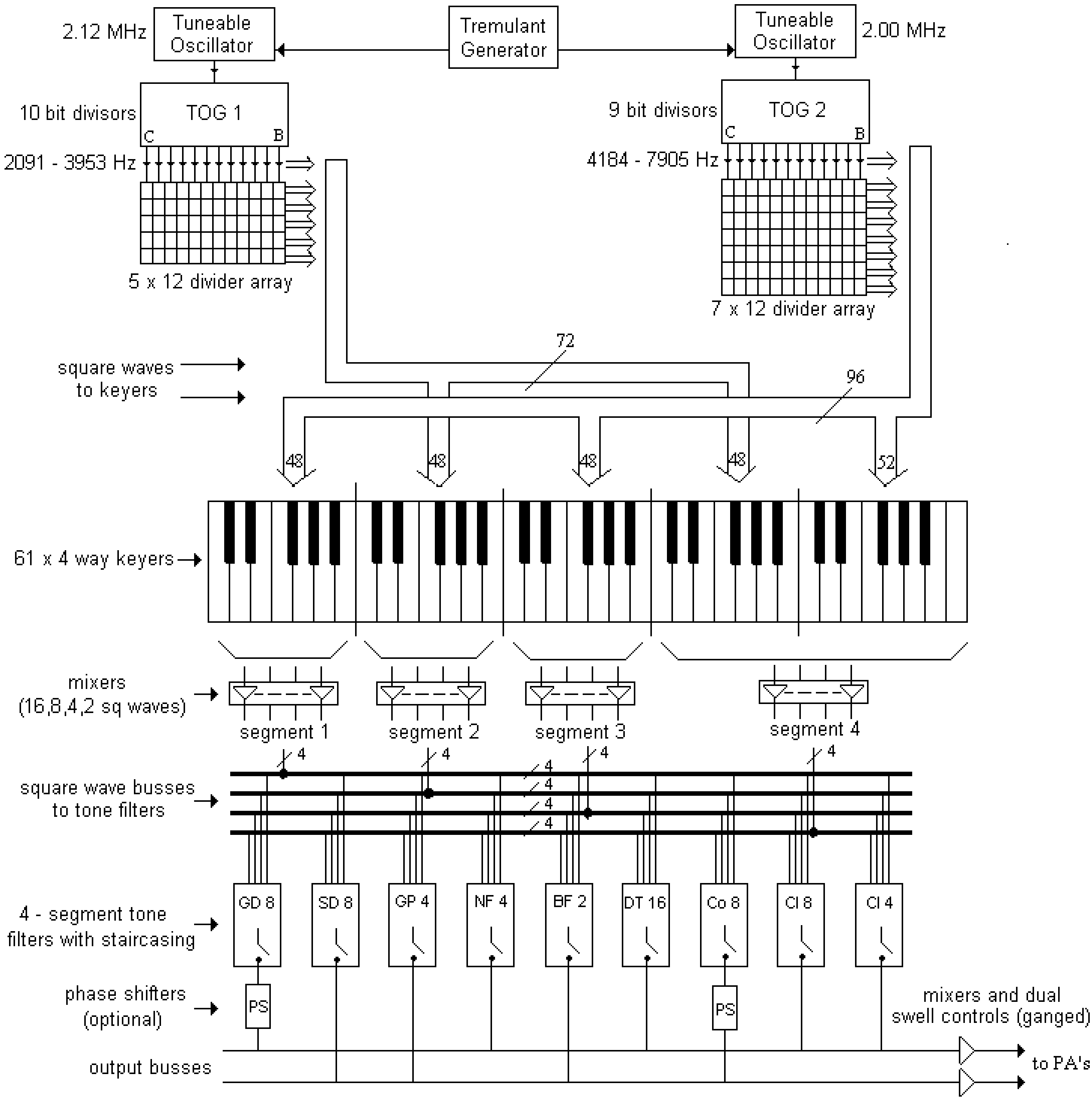


Figure 1. Swell department block diagram (divider system)

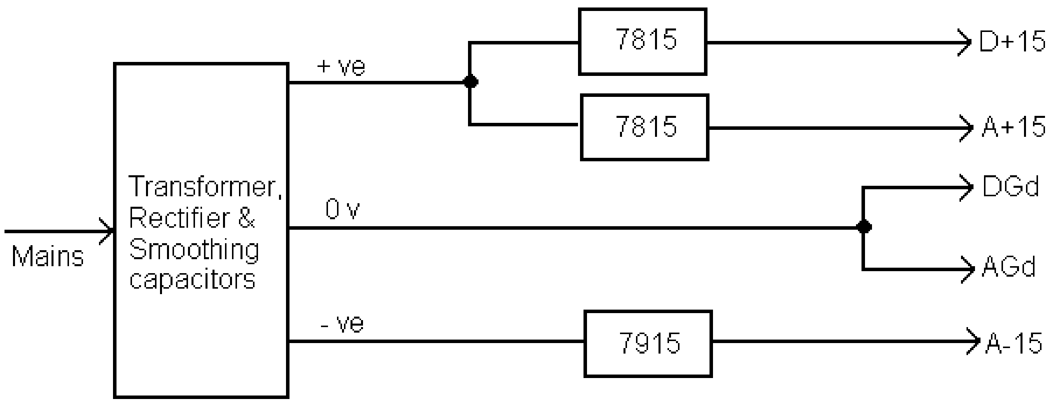


Figure 2. Suggested PSU arrangements

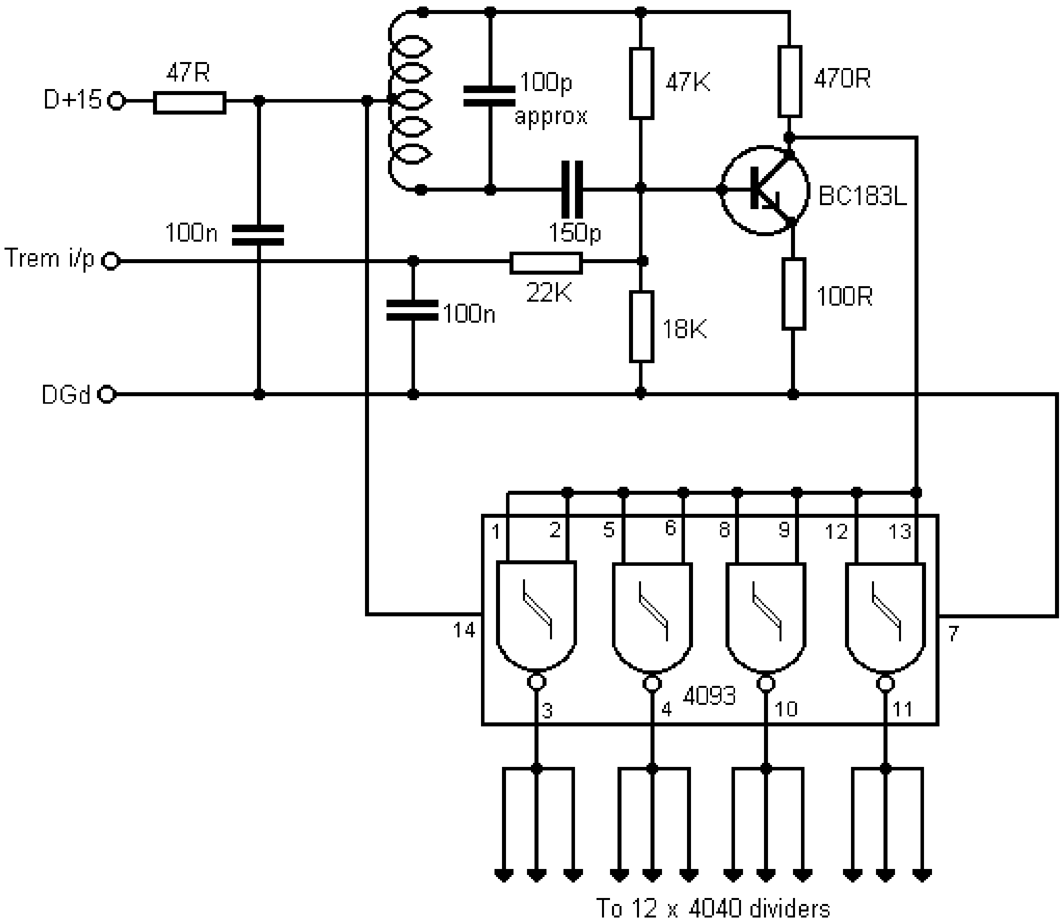


Figure 3. Oscillator and buffer circuit

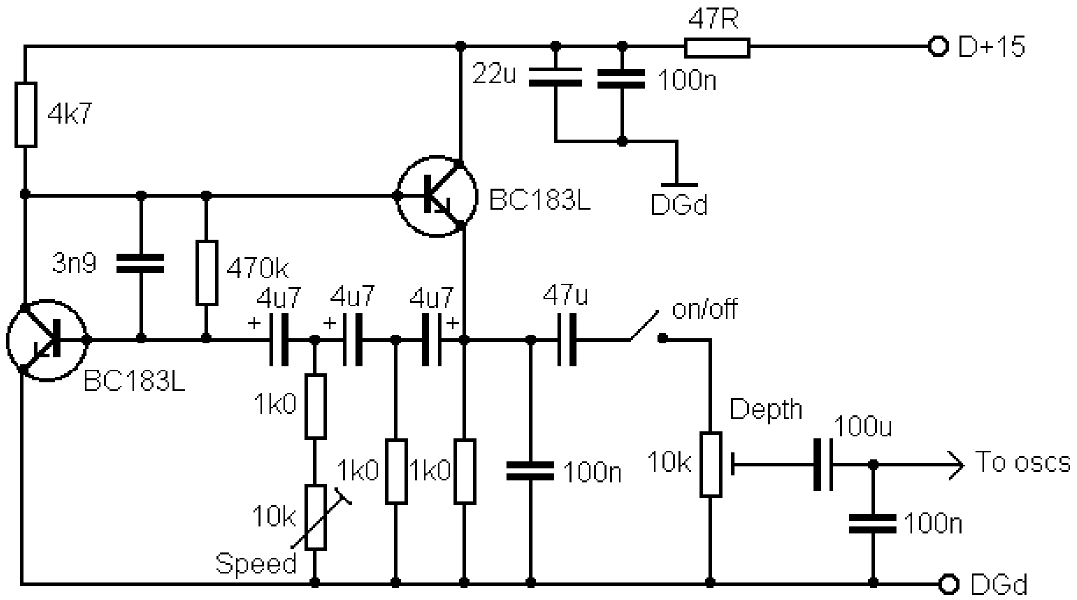
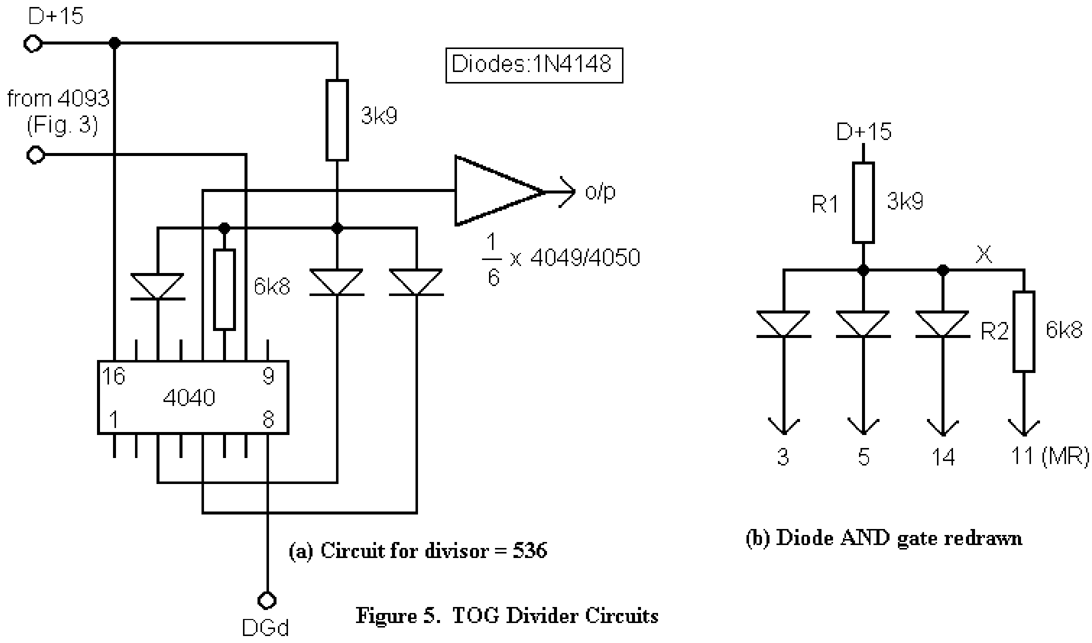


Figure 4. Tremulant oscillator



(a) Circuit for divisor = 536

(b) Diode AND gate redrawn

Figure 5. TOG Divider Circuits

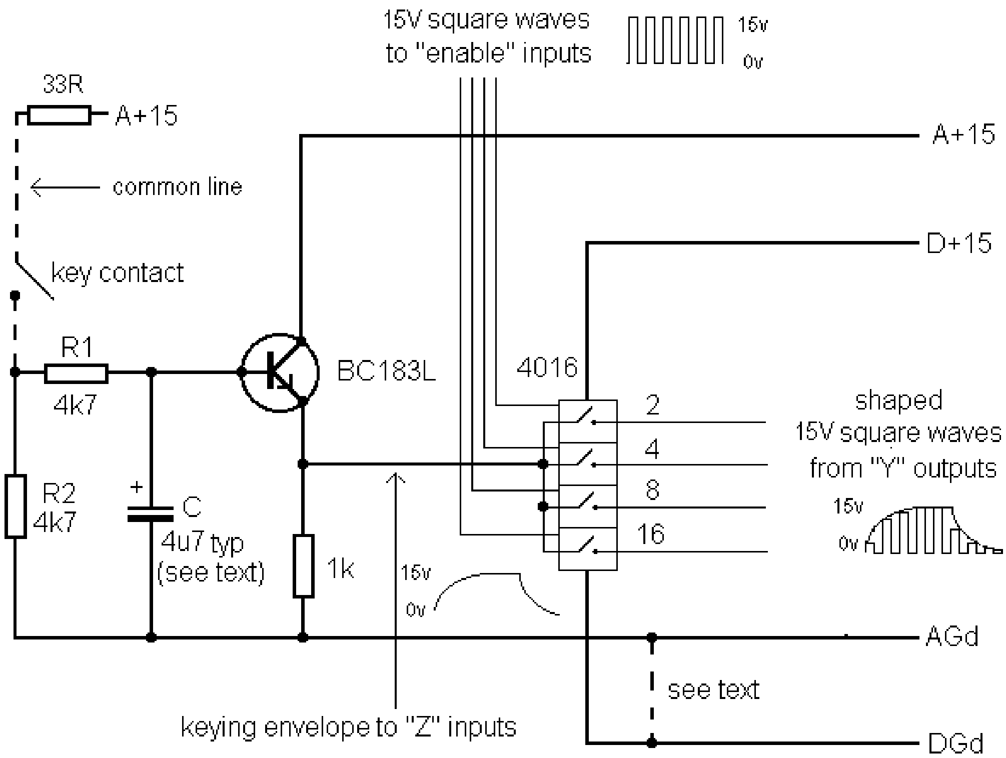


Figure 6. Basic keying circuit

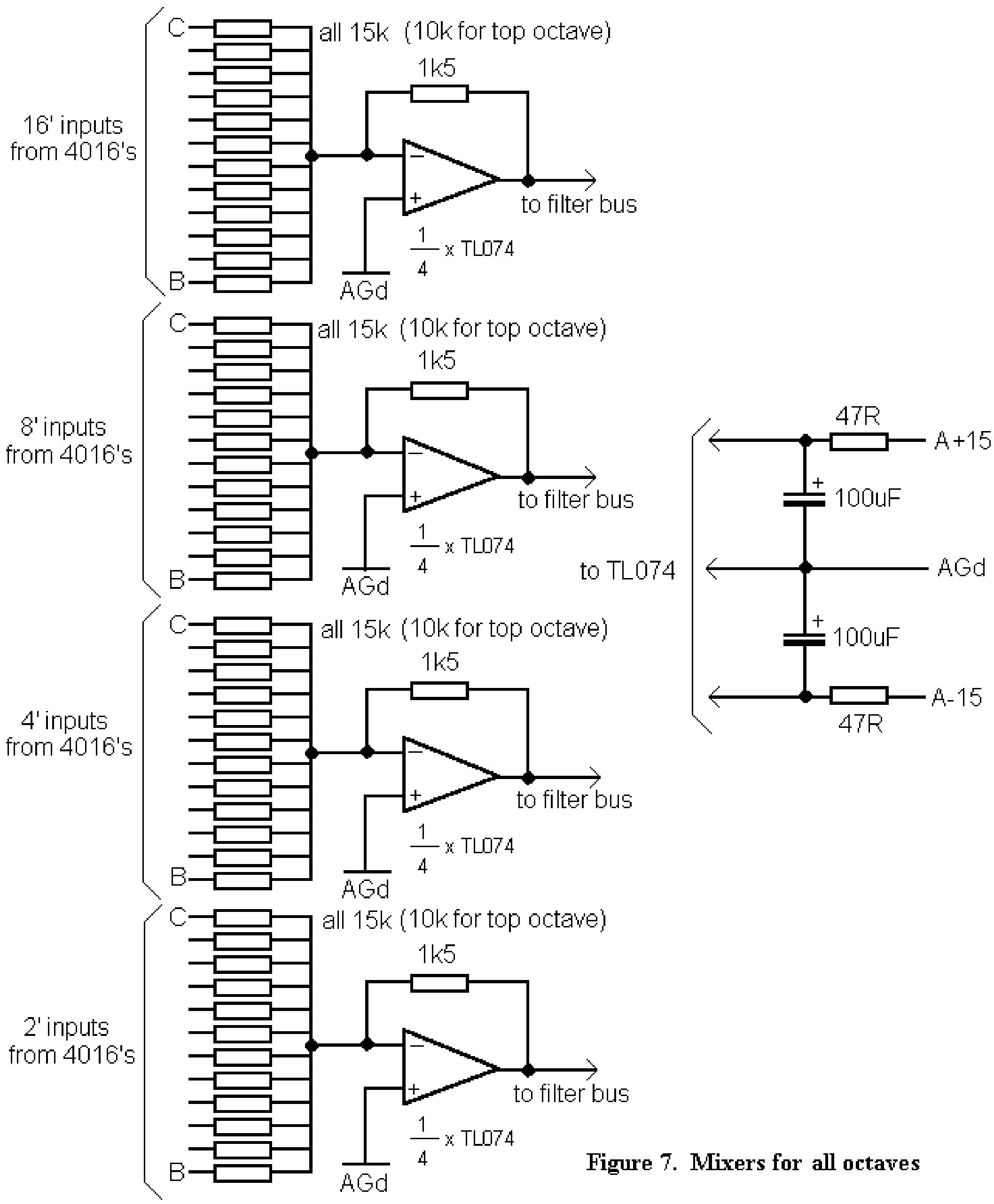


Figure 7. Mixers for all octaves

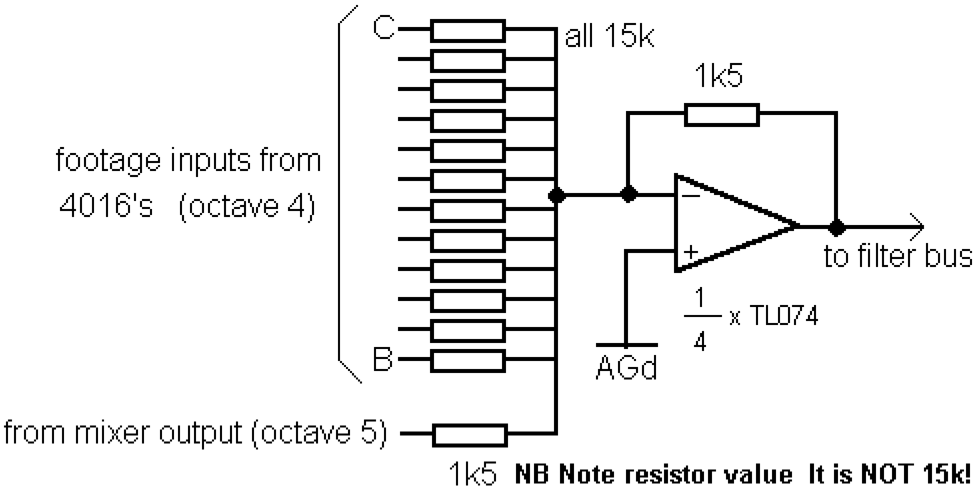


Figure 8. Combining the outputs of octaves 4 and 5 (only one footage shown)

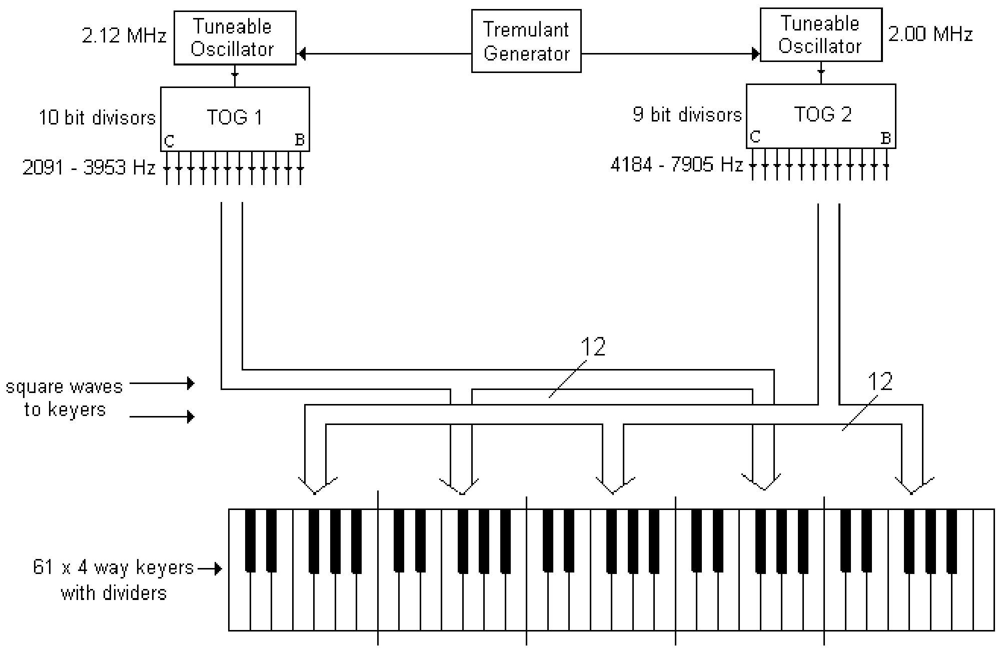


Figure 9. Alternative frequency distribution arrangement.

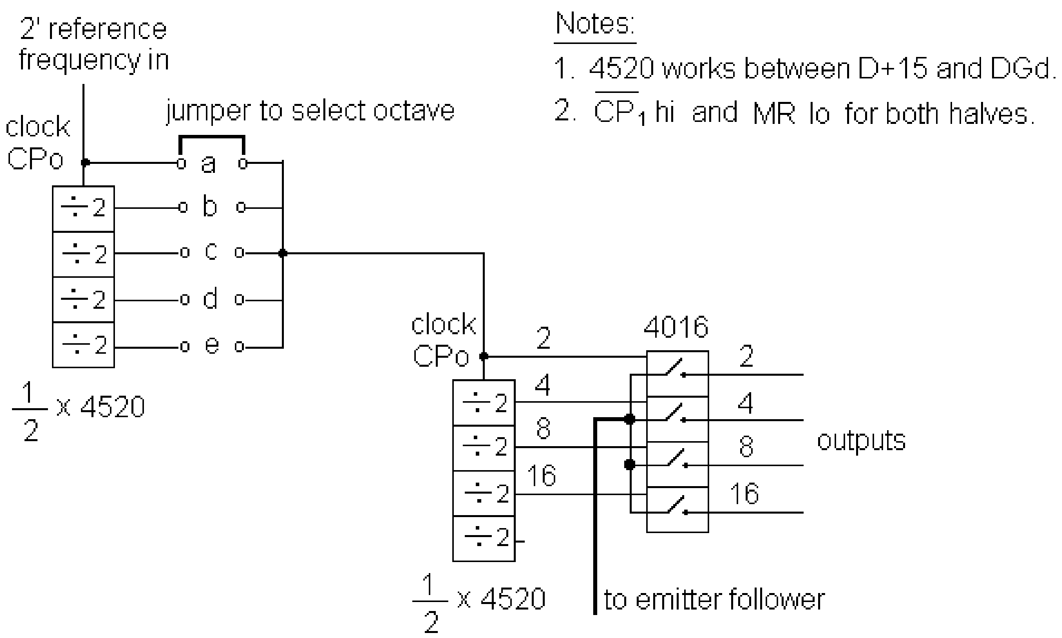


Figure 10. Keyers with frequency division.

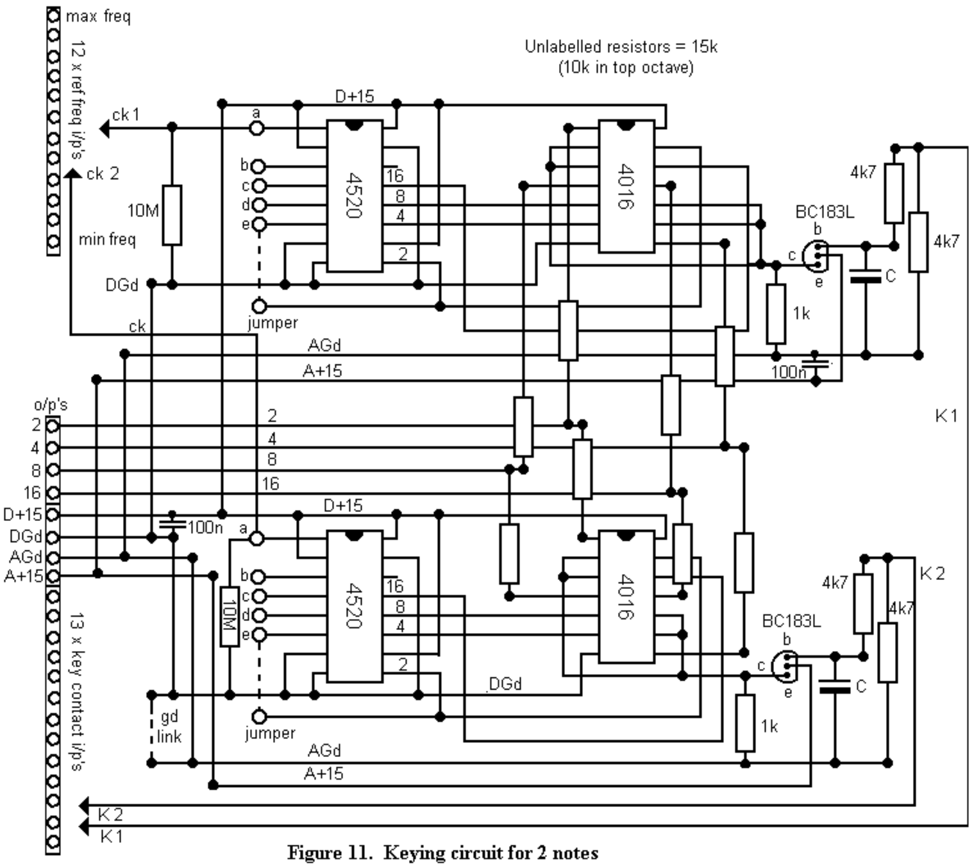


Figure 11. Keying circuit for 2 notes

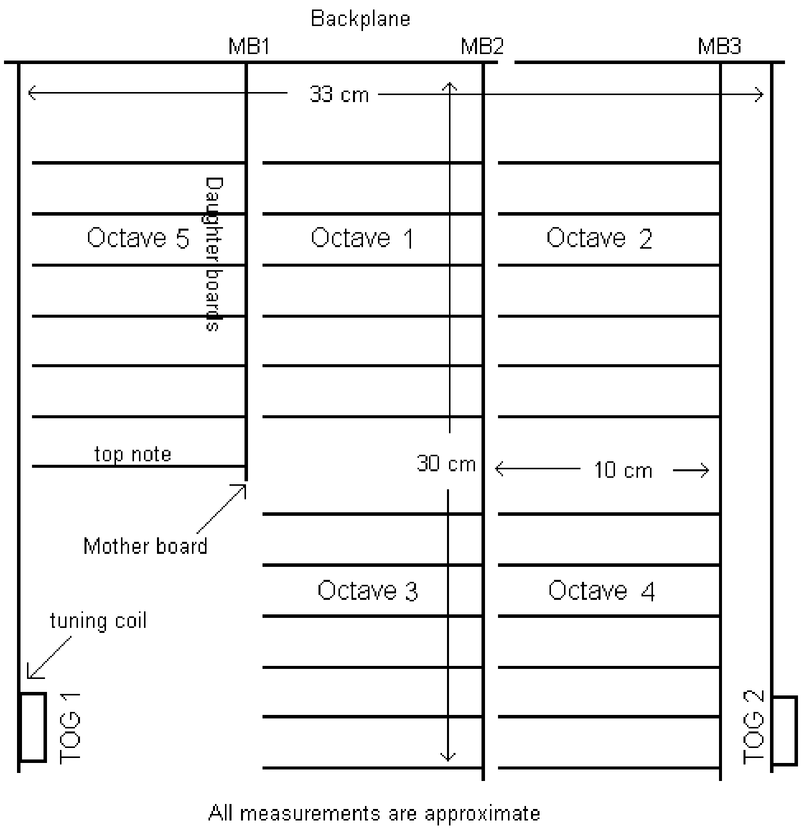


Figure 12. Plan view of keying system

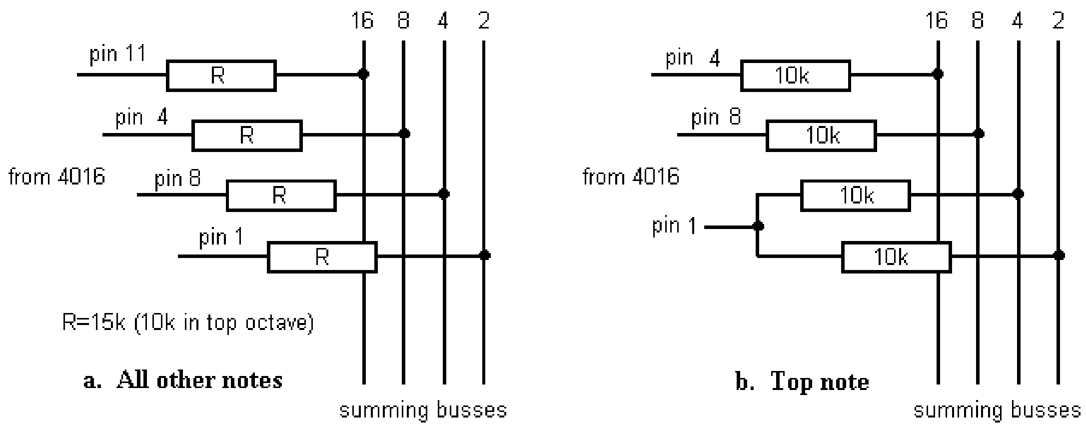


Figure 13. Mixer connections for top note

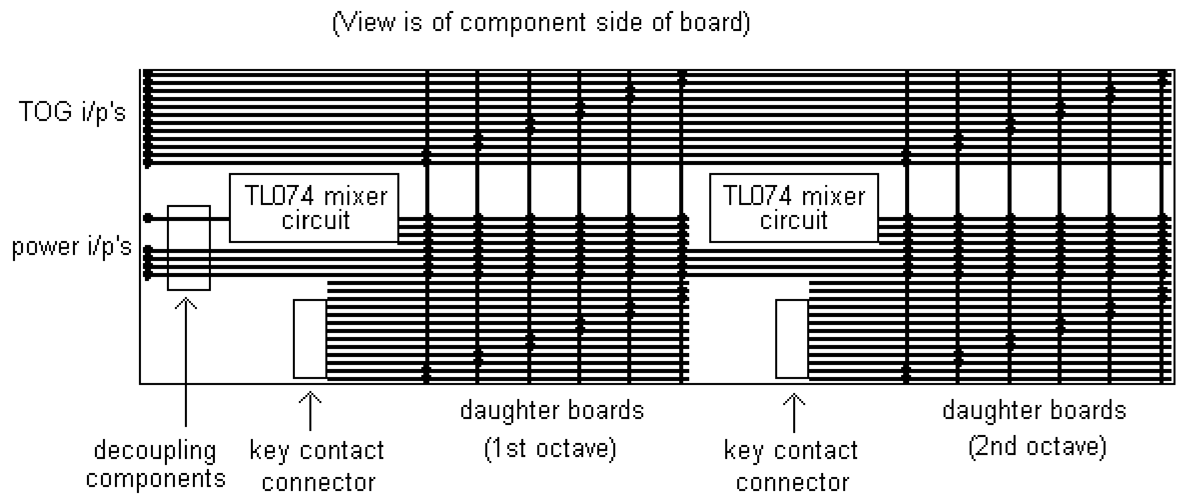


Figure 14. Mother/daughter board arrangement

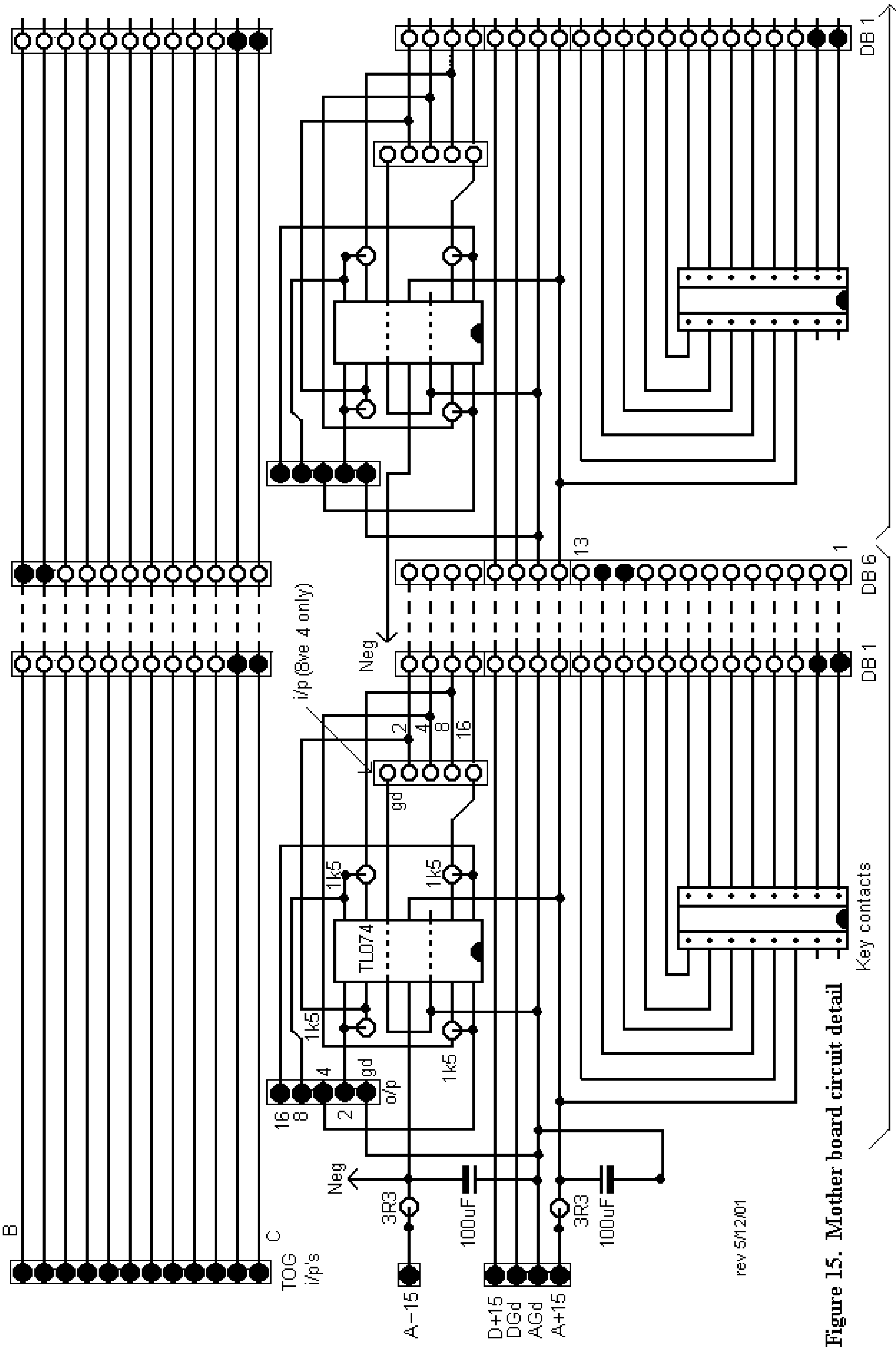


Figure 15. Mother board circuit detail

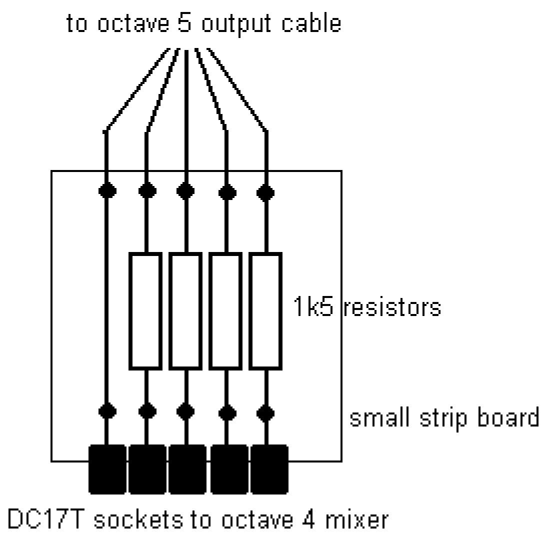


Figure 16. Octaves 4 & 5 connector arrangement.

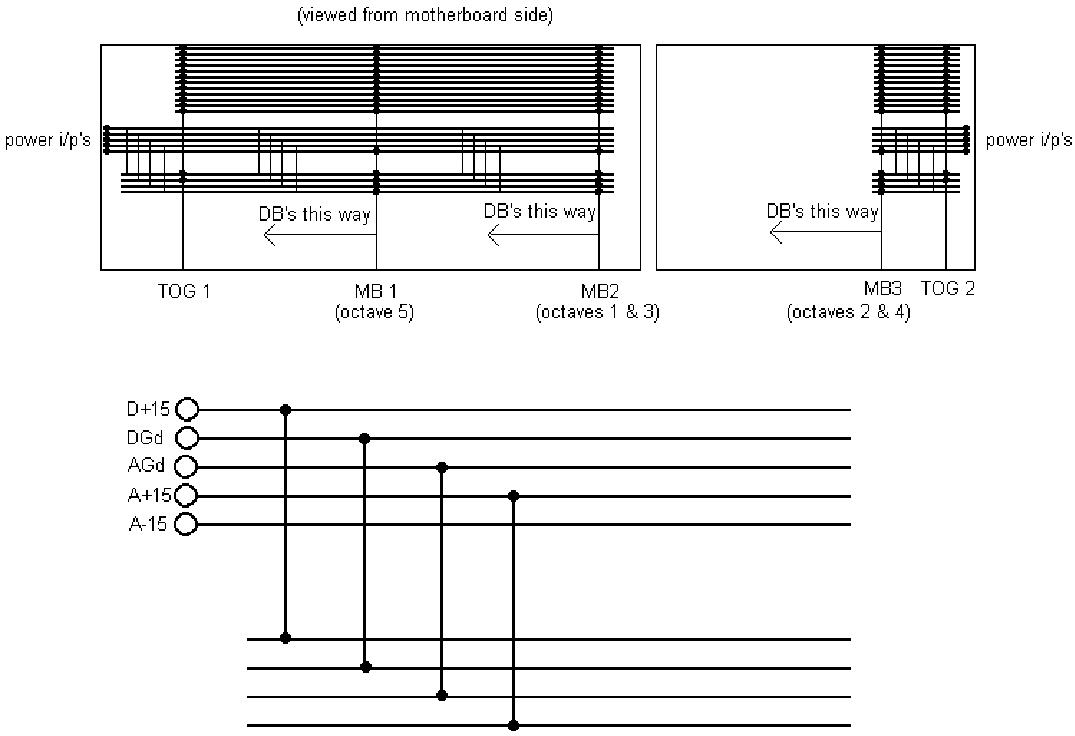


Figure 17. Backplane arrangement

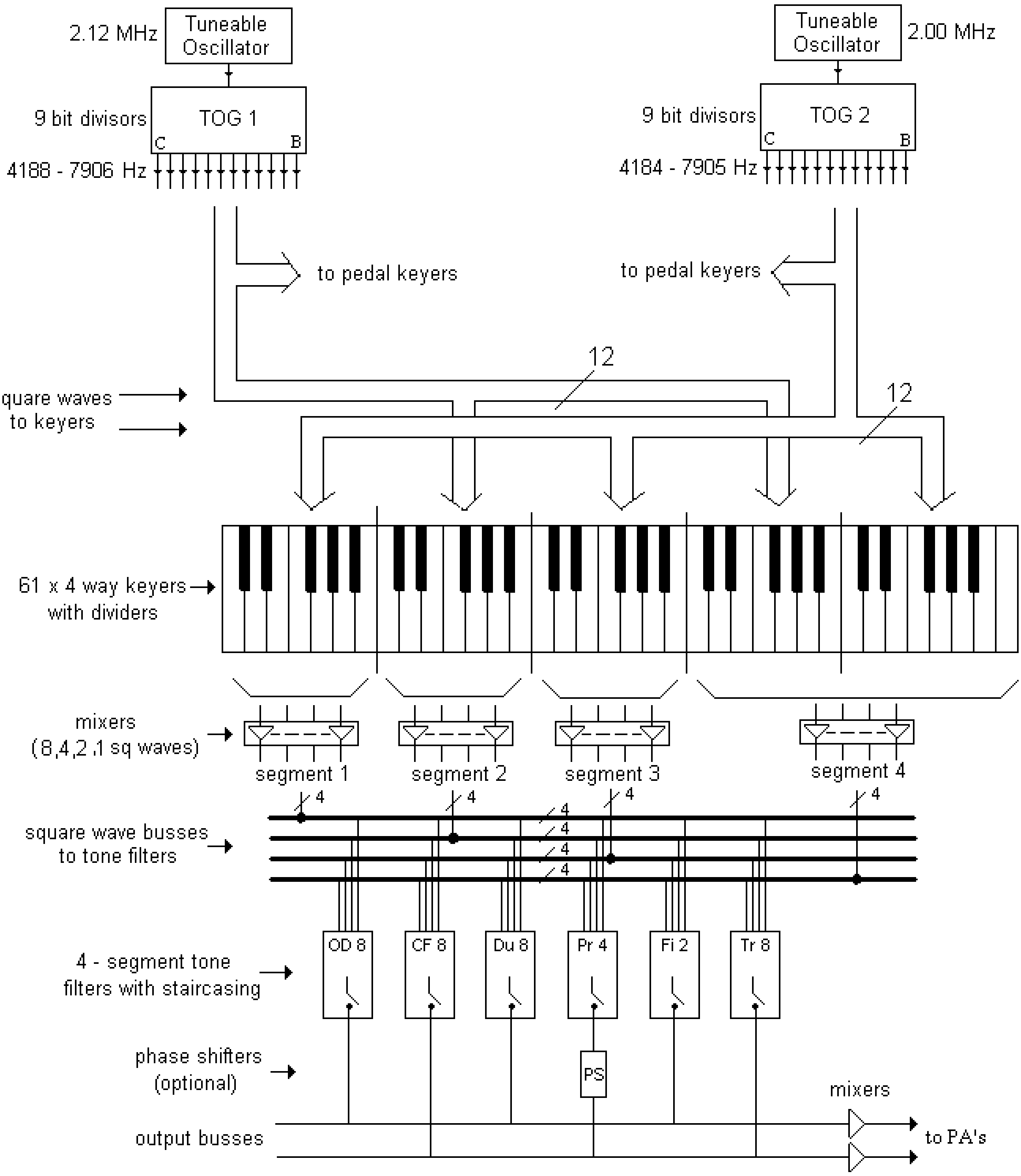


Figure 18. Great department block diagram (divider system)

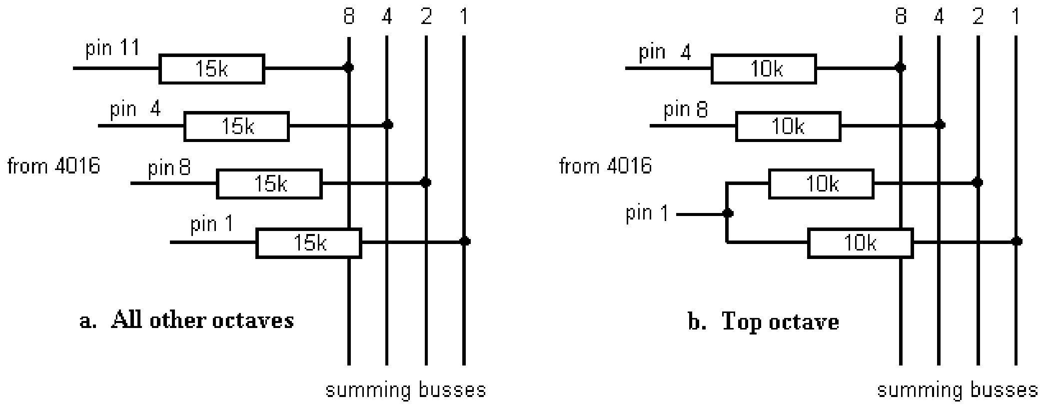


Figure 19. Mixer connections for top octave (great)

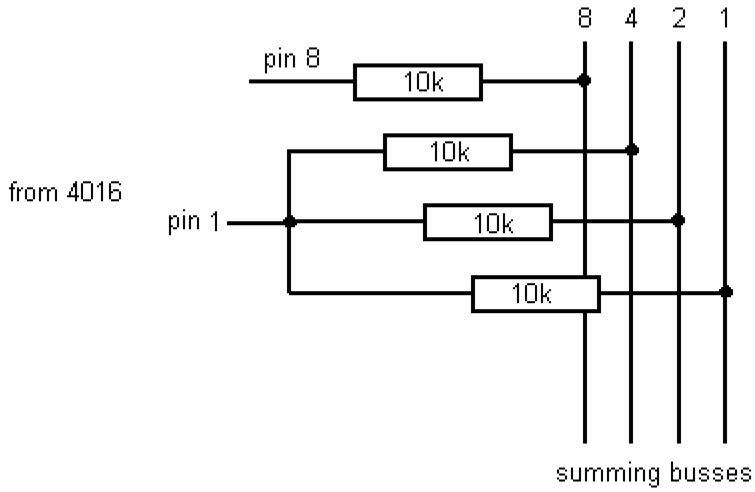


Figure 20. Mixer connections for top note (great)

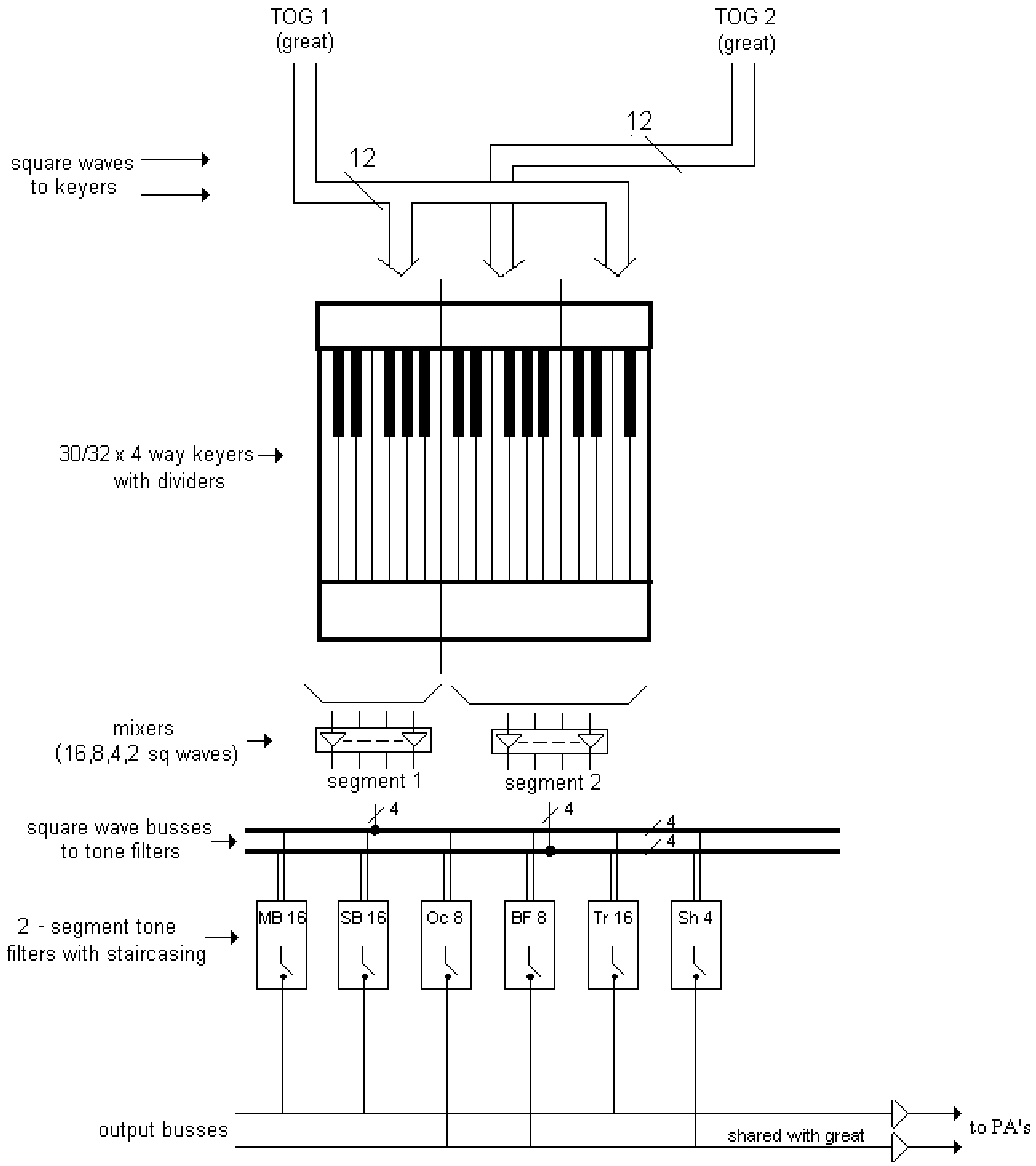


Figure 21. Pedal department block diagram (divider system)

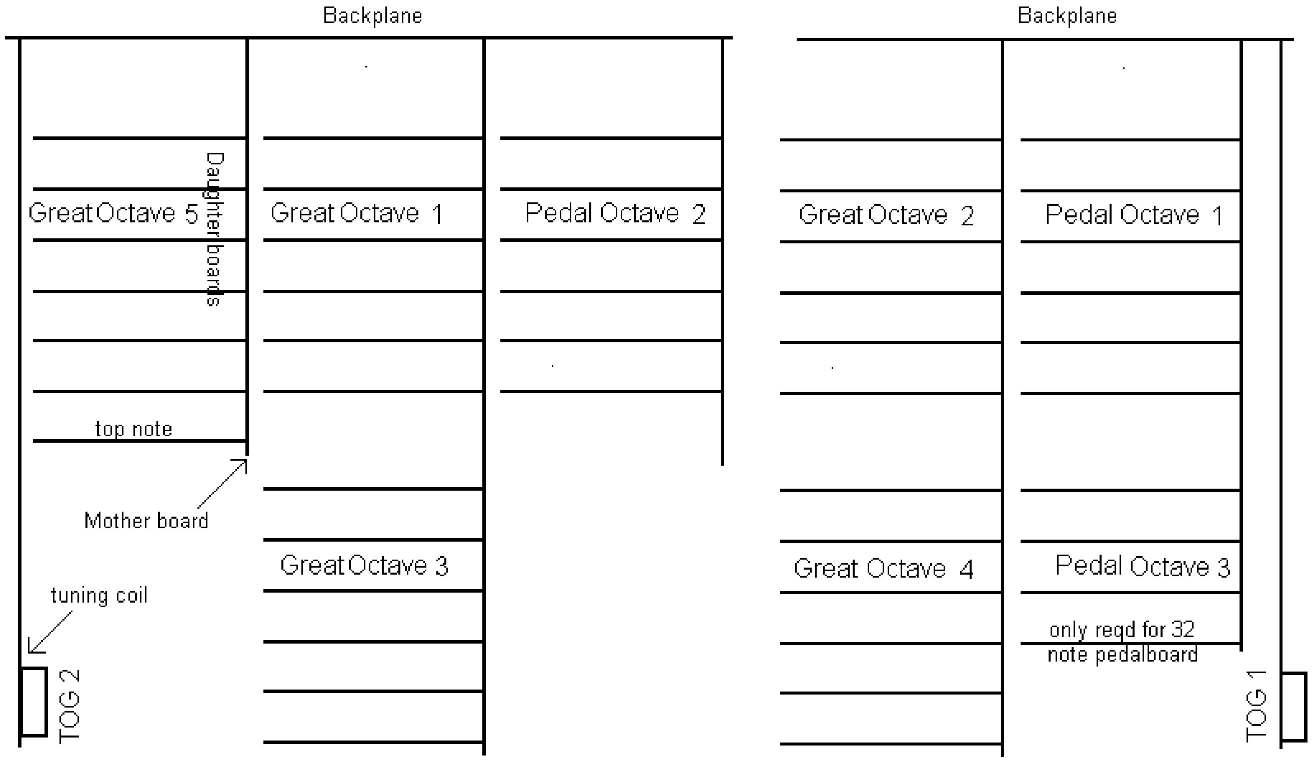


Figure 22. Plan view of great & pedal keying system

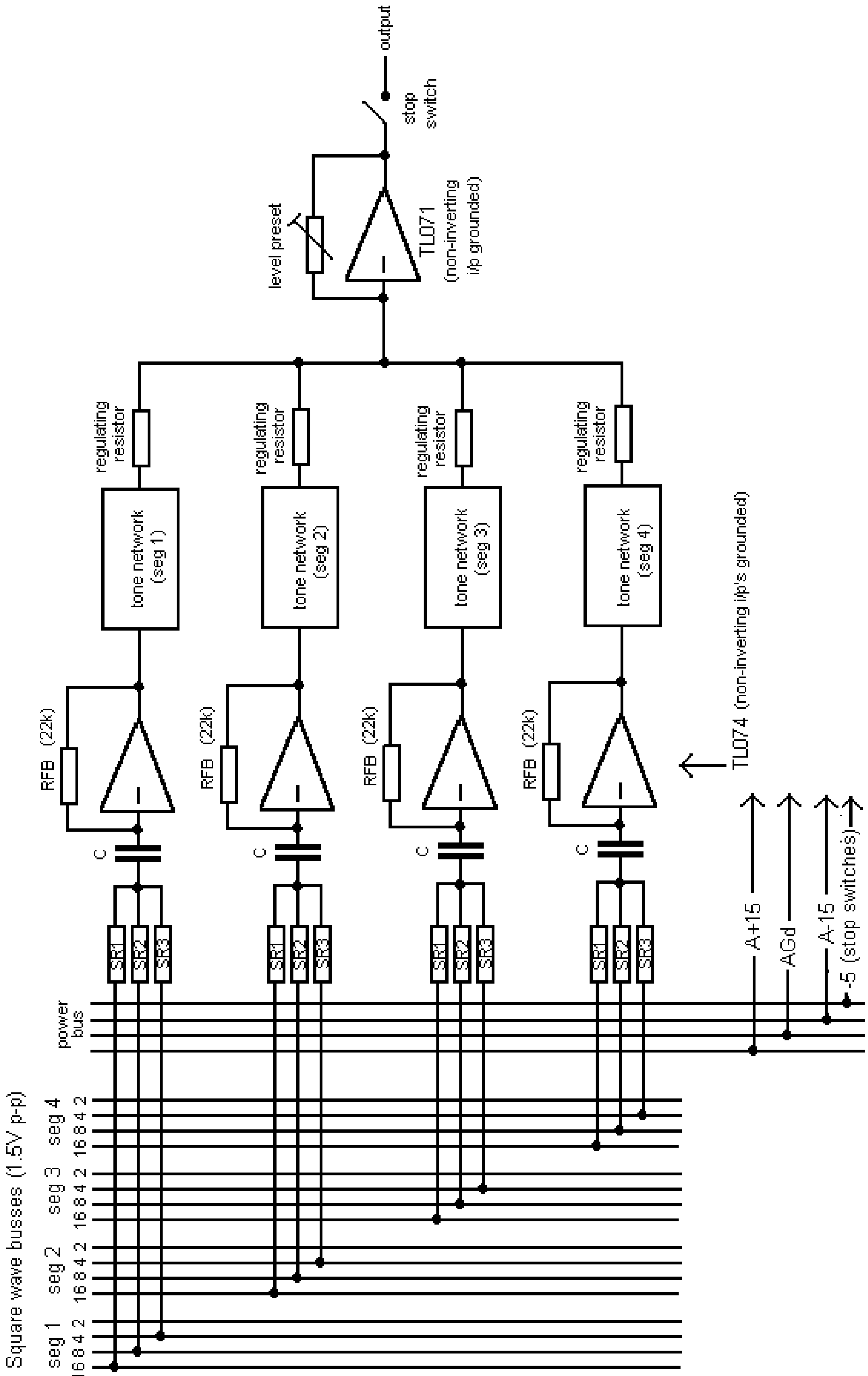


Figure 23. Tone filters - general arrangement for one stop

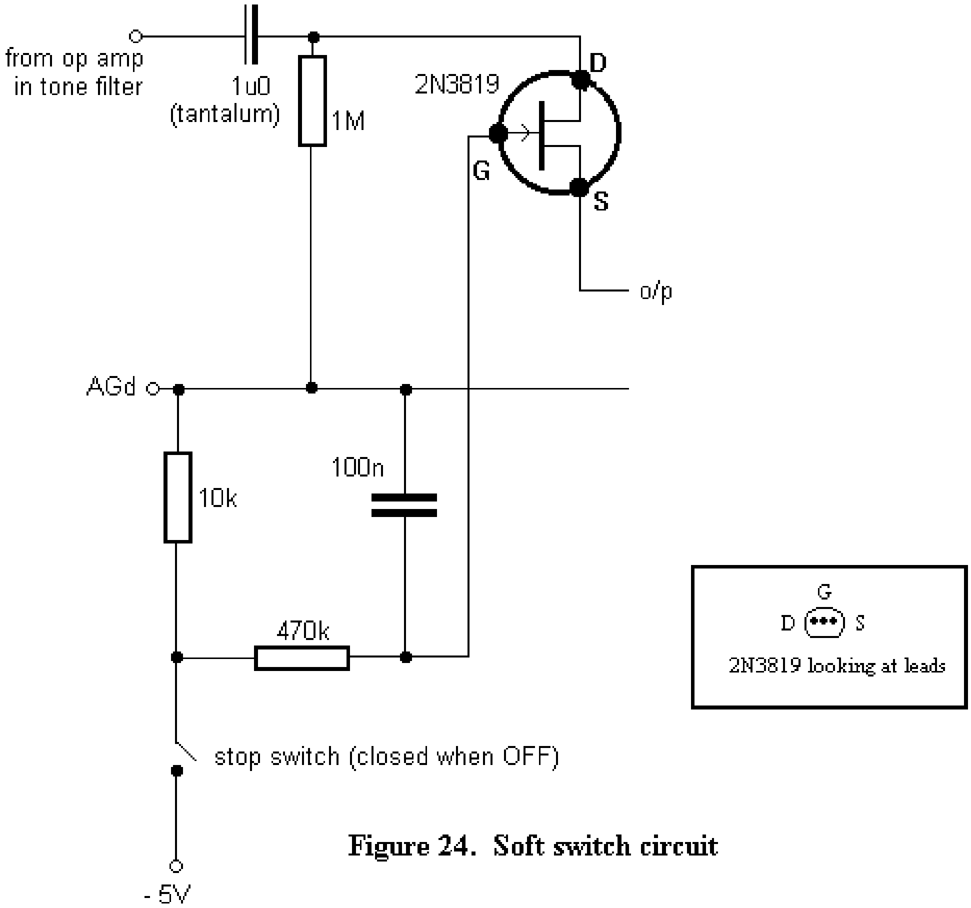
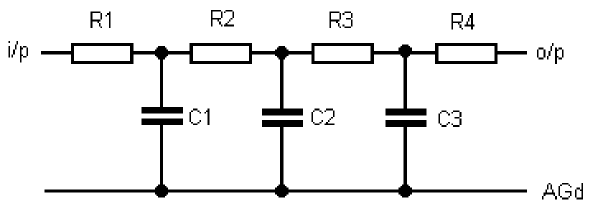
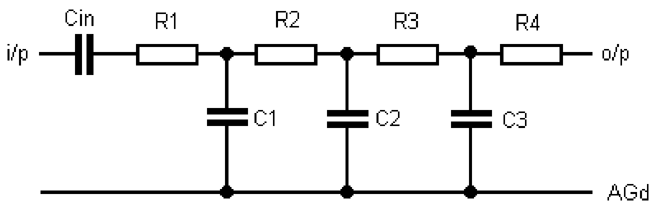


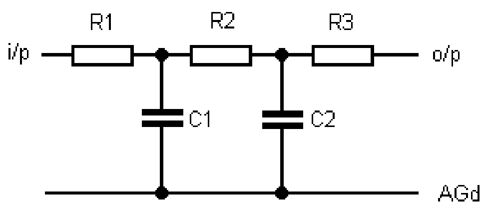
Figure 24. Soft switch circuit



(a) Geigen Diapason, Nason Flute, Claribel Flute, Dulciana, Principal, Major Bass, Sub Bass and Bass Flute

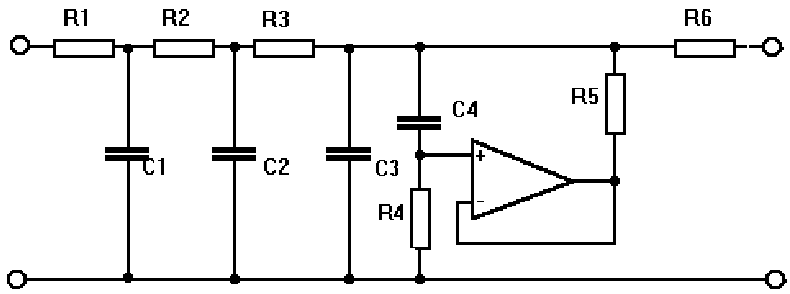


(b) Stopped Diapason and Block Flute

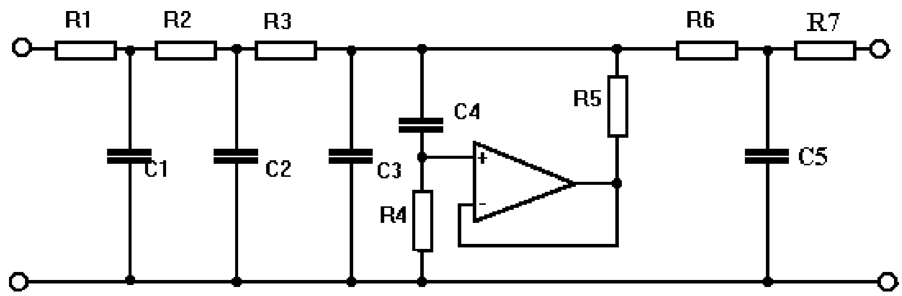


(c) Geigen Principal and Fifteenth

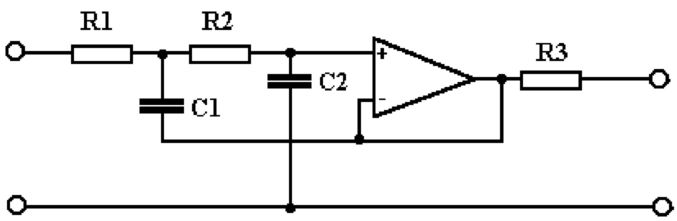
Figure 25. Tone filter networks



(a) Double Trumpet, Clarion and Trombone



(b) Cornopean, Clarinet, Trumpet and Schalmey



(c) Open Diapason

Figure 26. Tone filter networks

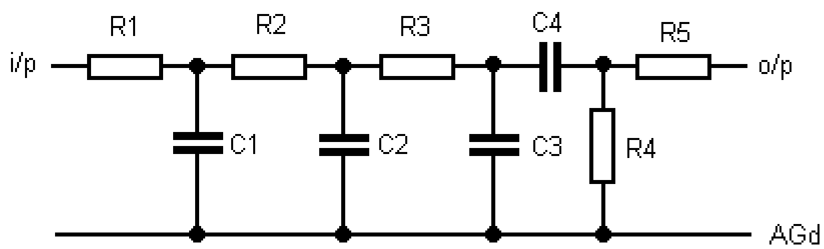


Figure 27. Tone filter network for Pedal Octave

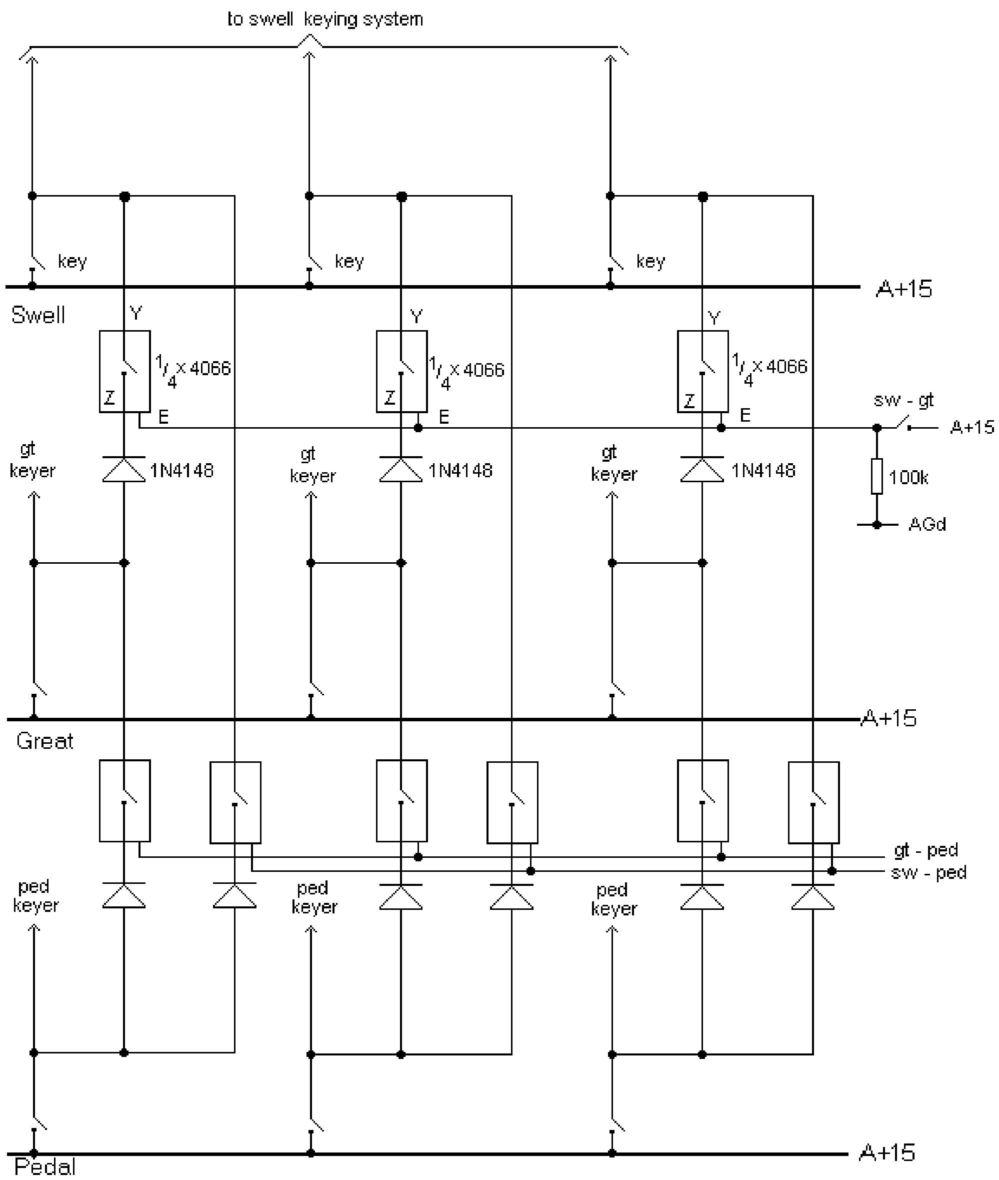


Figure 28. Coupling using one contact per key

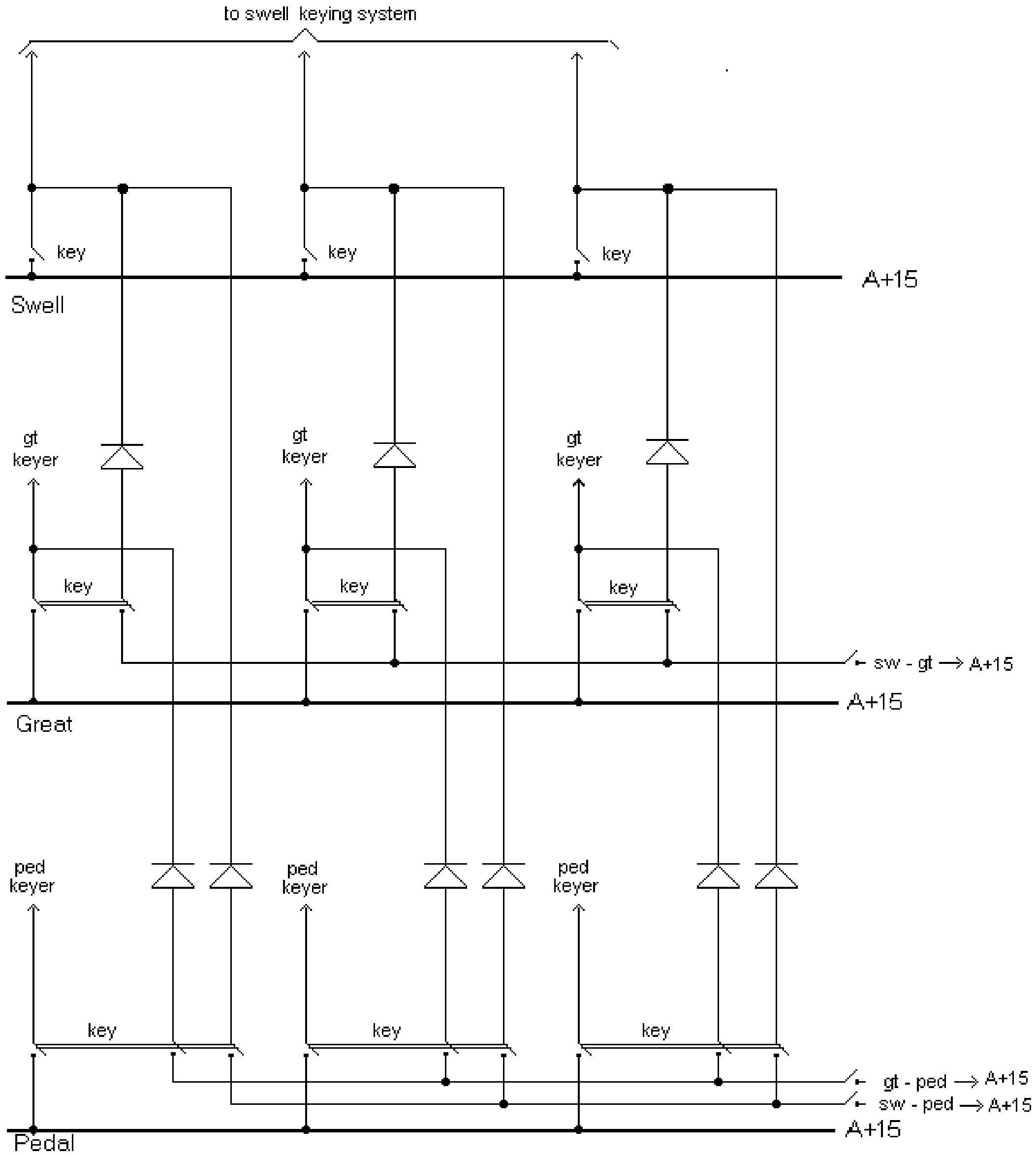


Figure 29. Coupling using multiple contacts per key

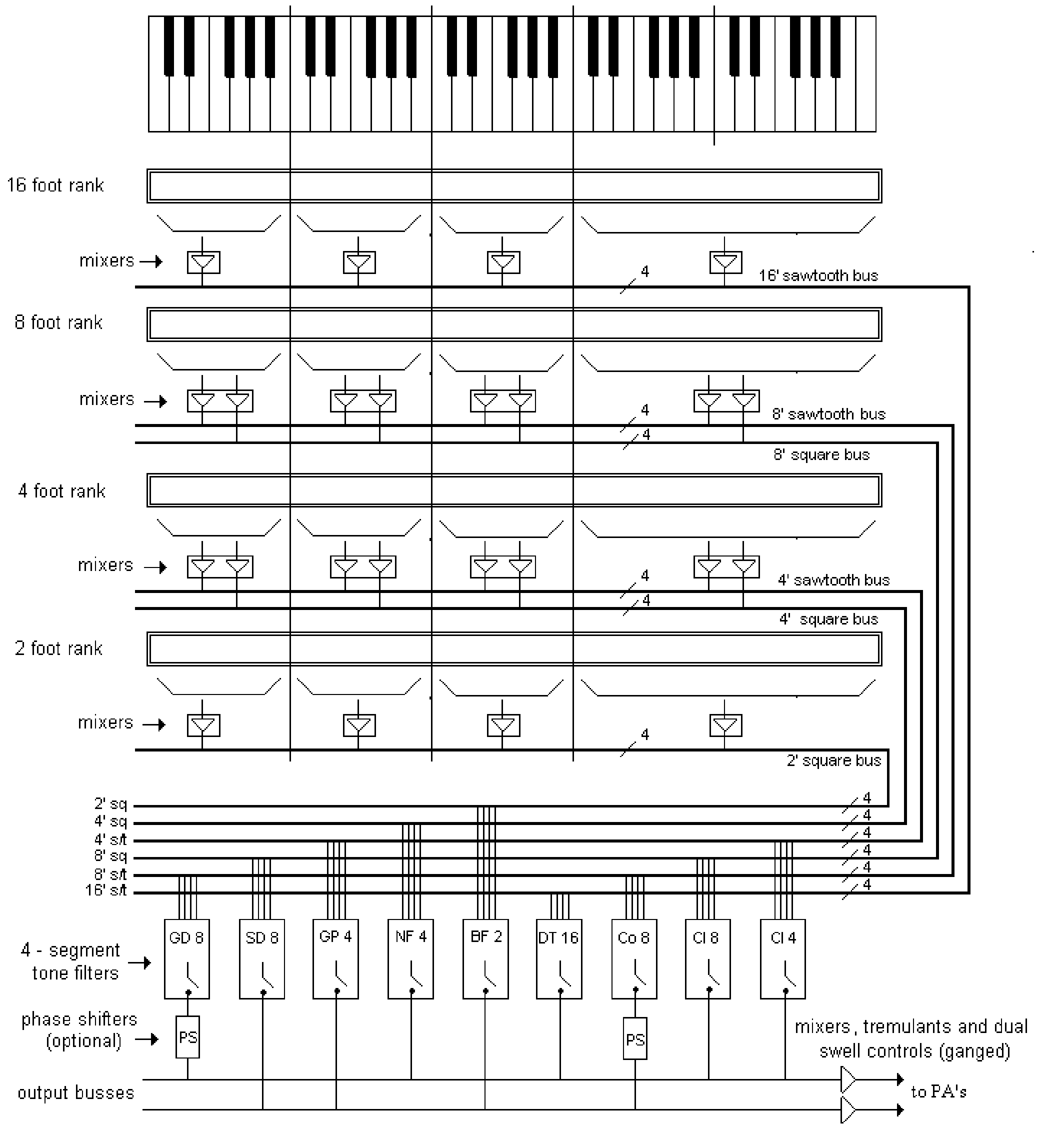


Figure 30. Swell department block diagram (free phase system)

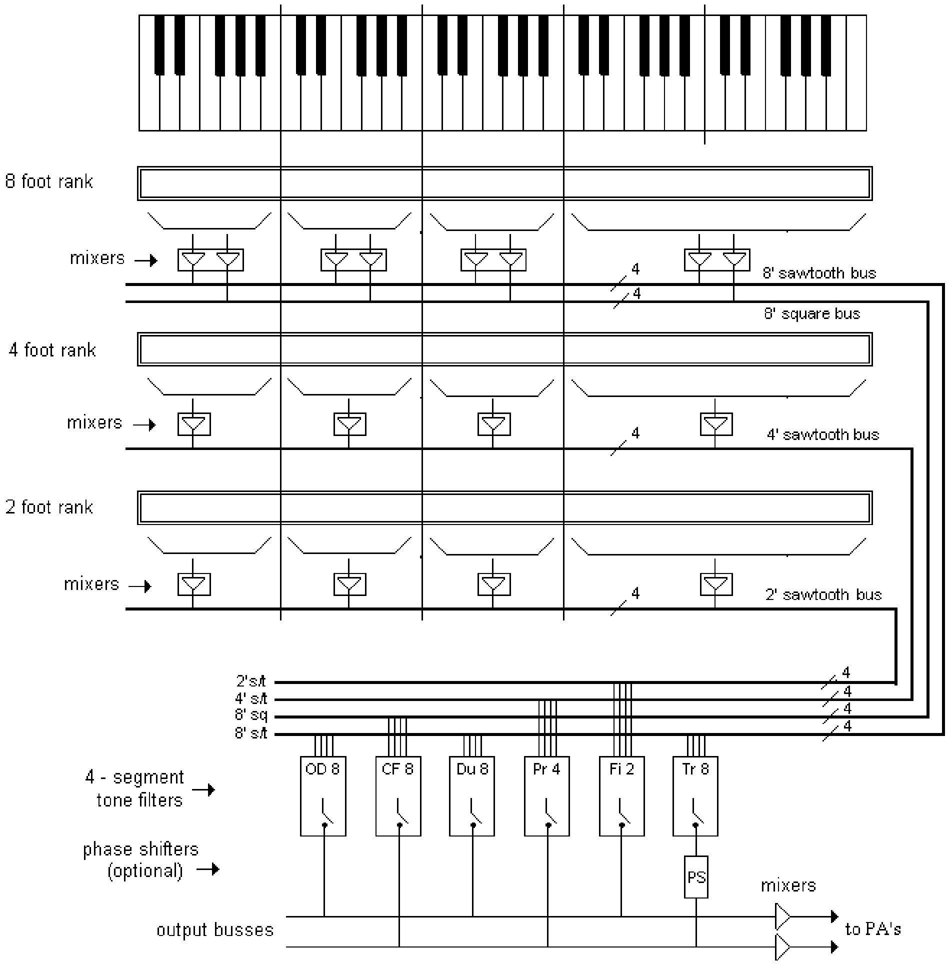


Figure 31. Great department block diagram (free phase system)

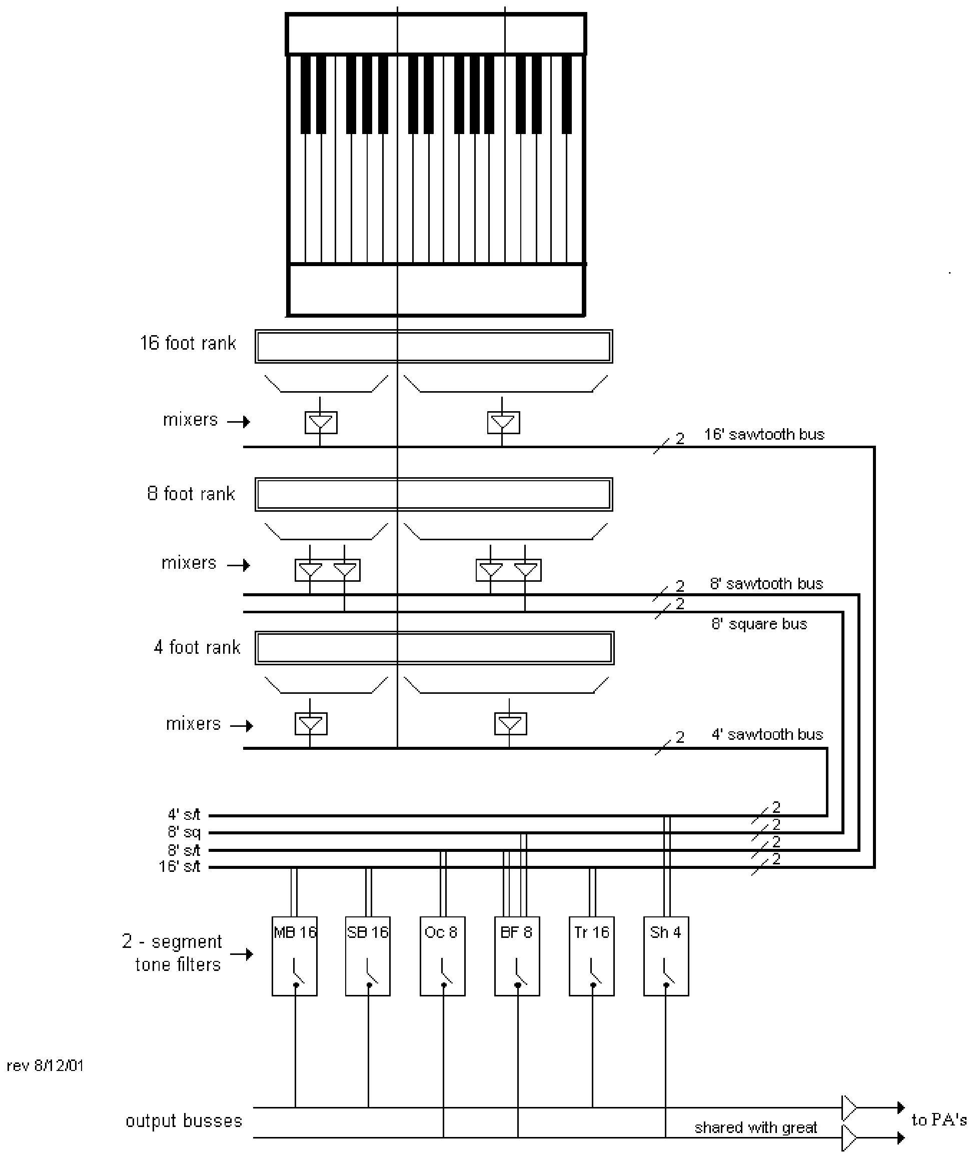


Figure 32. Pedal department block diagram (free phase system)

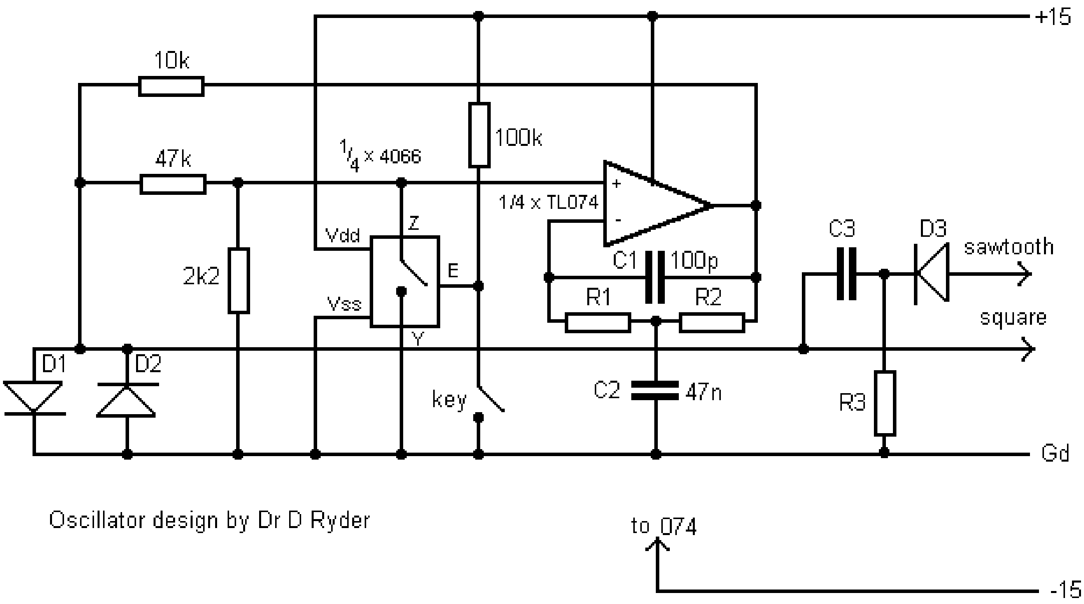


Figure 33. CFP oscillator giving sawtooth & square waves

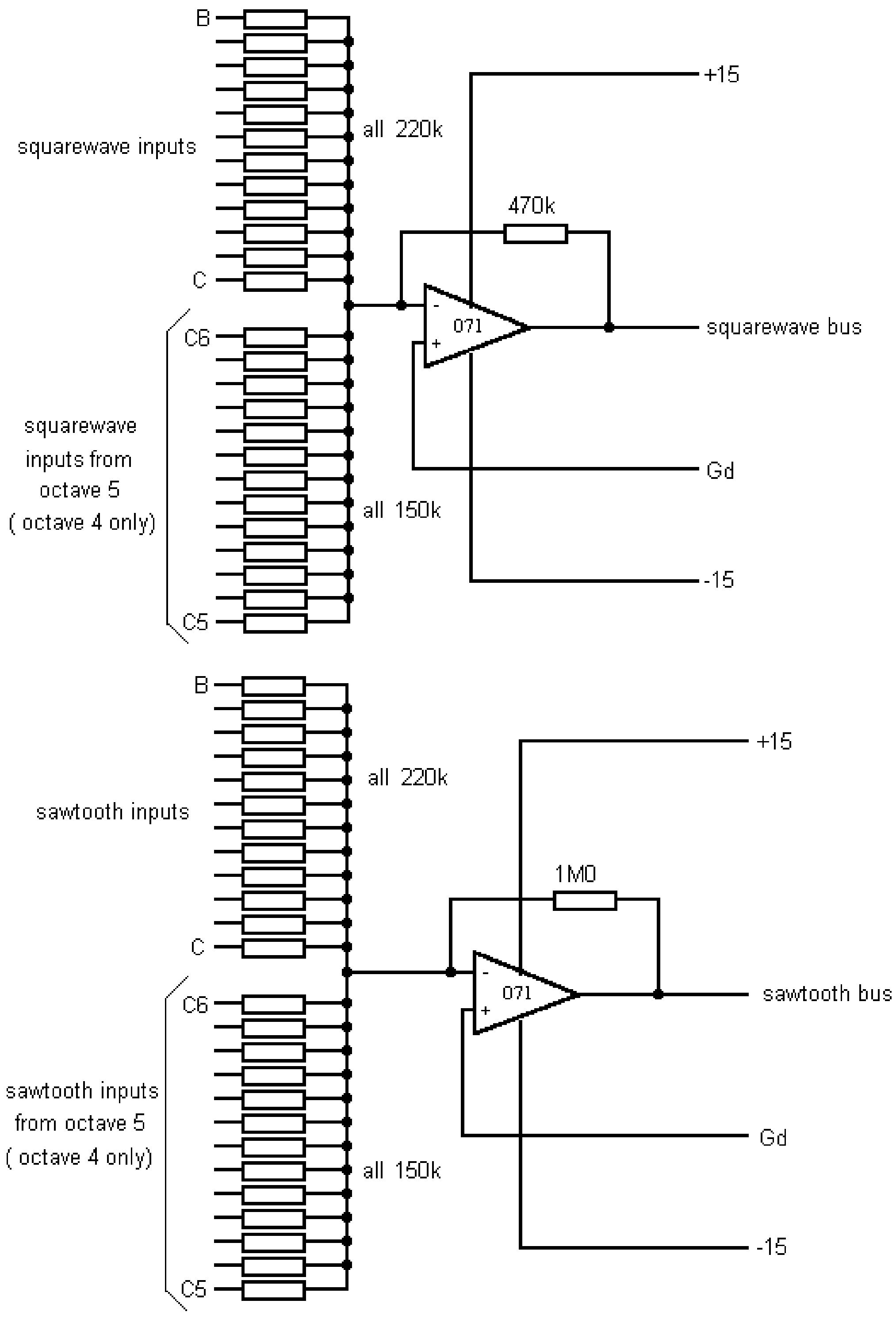


Figure 34. Segment mixers for free phase system (rev 01/10/02)

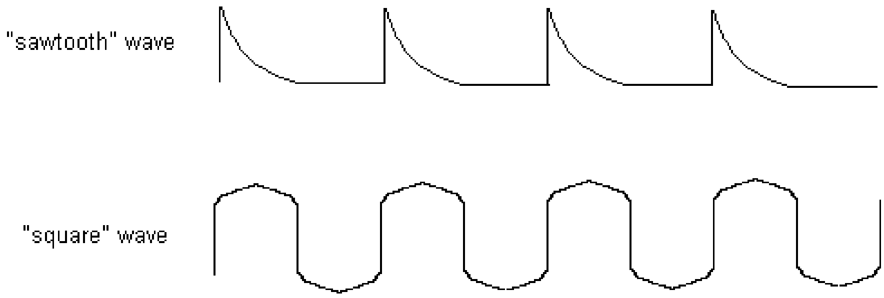


Figure 35. Waveforms produced by the CFP oscillators

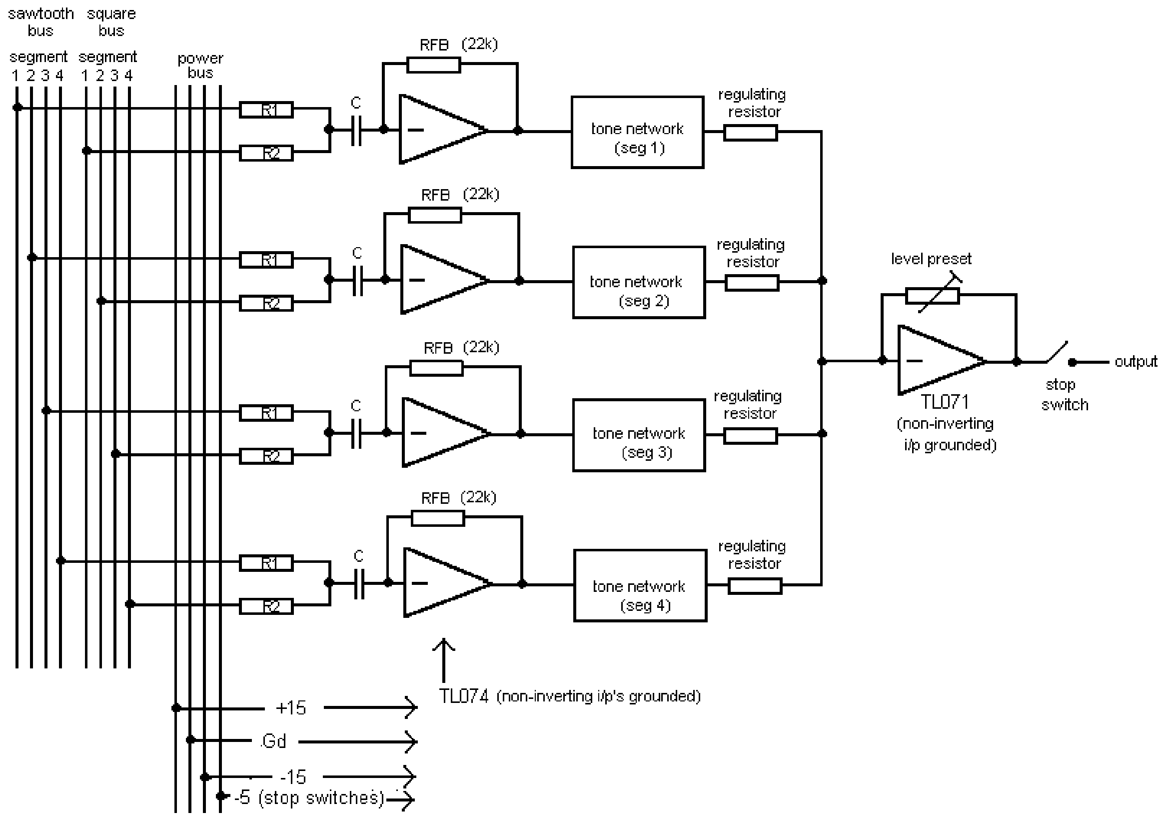


Figure 36. Tone filters - general arrangement for CFP oscillators (rev 01/10/02)

APPENDICES

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APPENDIX 1

The Reproduction of Very Low Frequencies

by Colin E. Pykett

(This updated article first appeared in EOM 118, November 1985)

The requirement to radiate acoustic energy at frequencies down to at least 30 Hz in a domestic environment is one which has caused continuing problems for those of us who have a penchant for organ music, either reproduced or played direct on an electronic instrument. Whilst proper reproduction of the Sydney Town Hall organ demands a house-rattling capability down to 8 Hz, I shall be reasonable in this article by setting 30 Hz as my lower limit (corresponding approximately to the frequency radiated by a 16 foot open organ pipe - bottom C on the pedals). Loudspeaker system design in this frequency region is complicated by the fact that not only is the response of the loudspeaker falling off (typically by at least 12 dB per octave, sometimes more) but so also is that of the ear. Thus subjectively satisfying bass reproduction may involve acoustic powers of several watts compared with milliwatts elsewhere in the spectrum (Fig. 1). Whilst the provision of electrical power is no longer a problem, the satisfactory conversion of it into acoustic power is still not easily solved.

> The curves in Figure 1 (a) are the Churcher-King contours of equal loudness. Note firstly that at all sound intensity levels the apparent loudness of low frequencies is much less than of frequencies around 3 kHz at which the ear is most sensitive. Secondly as the absolute intensity reduces this discrepancy becomes more marked. This reflects the situation in domestic rooms where your neighbours would not usually wish you to radiate powers around the 100 **acoustic** watts which characterise a cathedral organ going flat out! Thirdly an added effect is the fall-off in bass response of most loudspeaker systems in the very low frequency region, illustrated by the curve in Figure 1 (b) for an expensive design of moderate size.
>
> Taken together these curves illustrate the magnitude of the problem we are up against in trying to get subjectively satisfactory bass response at comfortable listening levels in the home: a difference of up to 70 dB in sound pressure level between 30 Hz and 3 kHz may conceivably be required. At least we might try to improve the loudspeaker response as a first step.

This article attempts to derive a clear view of the physics of the situation with a promise of no maths, jargon or other mysticism. Thus I hope you will relax and read what follows with your feet up on the mantelpiece. Having established the physical picture, some common loudspeaker designs are discussed followed by some observations based on personal experience. I hope this will open up a way ahead for those who feel they have yet to reach their low-frequency Nirvana

Myths and Legends

Loudspeakers, probably more than any other component in the reproducing chain, have unfortunately been the subject of misunderstanding, myth and downright nonsense even amongst those who should know better. For example, reputable journals have devoted space over many years to rubbish such as the “advantages” of grossly expensive speaker cable over much cheaper ordinary material, whilst ignoring the very real detrimental effect of the contact resistance of flimsy DIN connectors which are quite unsuited to the heavy currents involved. As a result of this pseudoscientific attitude a number of myths continue to circulate. Some are listed below; all are either untrue or only true within certain caveats:

- Horn loudspeakers can be relied upon to provide a better bass response than direct-radiating loudspeakers (those which face direct into the room).

- Horns radiate with higher efficiencies than direct radiators

- A small sealed enclosure (mistakenly called an "infinite baffle") can produce good bass if it is driven hard enough.

- Bass response is better from a loudspeaker in a corner of a room.

- There is no directivity at low frequencies, therefore it does not matter where the bass loudspeakers are placed in the room.

How many of these do you subscribe to?

A Physical Picture - Big is Beautiful

A rigorous or even semi-rigorous analysis of the loudspeaker and its enclosure treated as an electro-acoustic transducer system is not fundamentally difficult in a mathematical sense but it involves some long-winded algebra, and (worse) tends to obscure the simple picture of how a moving object in the air (a loudspeaker cone) causes acoustic energy to propagate away from it

Consider a normal moving-coil loudspeaker mounted behind an aperture in a thick piece of wood (a baffle) about 10 metres in diameter. This is a large object and it approximates to a true infinite baffle down to very low frequencies indeed. In spite of its inconvenient size, it is nevertheless a useful starting point for the discussion. Because of the low frequency assumption we can be confident that the loudspeaker cone moves to and fro as would a rigid piston in response to an impressed sine wave without breaking up into various regions moving in different directions, a phenomenon which occurs at higher frequencies. Consequently when the cone moves forward, air in front is compressed. That at the rear is rarefied, and the air pressure attempts to equalise itself by moving around the edges of the baffle. Since in this case the baffle is large, the time taken for these equalisation movements is also large (limited by the speed of sound, which is the fastest that pressure waves can travel). For a 10 metre baffle, the time taken for a pressure increase at the front of the cone to reach a corresponding point at the rear is about 30 msec. Since frequency is the reciprocal of time this corresponds to about 33 Hz, a fact which shows that below this frequency there will be time for the air pressure to substantially equalise as the cone moves, thereby reducing the amplitude of the pressure changes which propagate to the listener who interprets them as sound. Thus low frequency reproduction demands a large baffle, an inescapable fact, though a more detailed analysis backed up by practical experience shows that 10 metres is a bit excessive for 30 Hz and 3 metres would be quite good. Even this, though, is domestically inconvenient.

We have not yet finished deriving our simple physical picture. If baffle size was all that mattered for bass reproduction then it might be imagined that a 2½ inch tweeter ought to be as good as an 18 inch woofer. This is plainly not so, though the influence of cone area on bass reproduction is less well understood than the effect of a baffle. It is loudspeaker size that we shall now explore.

Assume there is a truly infinite baffle (so that pressure equalisation effects do not occur) and imagine what happens to the air in front of the cone as it moves forward. Initially, the cone compresses a cylindrical column of air in front of itself. The local pressure enhancement within the column is dissipated by air movement from inside to outside the column, leading to propagation of a disturbance which causes the sensation of sound. Intuitively we may see that the wider the column (i.e. the larger the loudspeaker) then the longer the equalisation process will take, because a high pressure area in the middle of a fat column has further to move before it dissipates than if it was in the middle of a thin column. This dissipation time governs the time for which outward sound propagation from the loudspeaker will occur in response to a cone movement and hence (if you think about it) the amplitude of the disturbance at a given distance from the speaker. Thinking further, it is possible to deduce that for efficient radiation of a particular frequency, the dimensions of the loudspeaker cone should be some significant fraction of a wavelength. If it is not then pressure dissipation, at the speed of sound, will take place in too short a time for the frequency being radiated. This is a fundamental requirement for all structures, whether acoustic or electromagnetic, which have to launch a disturbance into the environment efficiently. In the case of a loudspeaker a diameter corresponding to, say, half a wavelength at 30 Hz would be about 5 metres. Again, experience shows that this is not really required but, as for baffles, the bigger the better. Certainly an aperture one tenth of this, corresponding to an 18 inch loudspeaker, is getting a bit marginal.

Therefore we have a second important truth: the size of the loudspeaker unit itself must be as large as possible to radiate low frequencies. It need not be a single loudspeaker - a number of smaller units wired in phase is just as effective as a single one of the same total cone area and may be a lot cheaper. The mathematics shows that nature unexpectedly favours our endeavours regarding loudspeaker size in that radiating efficiency at low frequencies is proportional to the **square** of the cone area. Thus two loudspeakers close together are four times as effective as one of the same size. Incredibly, this point seems to be scarcely known.

Resonant Frequency

It is accepted wisdom that the resonant frequency of a loudspeaker should be as low as possible for good bass reproduction. This is true but at first sight it may not be obvious why - after all, any moving coil loudspeaker plainly has a response down to "DC" since a battery connected to the terminals will cause the cone to move to a new position and stay there until the voltage is removed (or the voice coil burns out).

The resonant frequency is principally determined by the mass of the moving elements (cone, etc) and by the stiffness of the suspension system, which incorporates a form of spring. Low stiffness and high mass contribute to the desirable low resonant frequency. Below resonance, the loudspeaker frequency response is dominated by stiffness. We can see this in the above case where the cone moved in response to an applied voltage step. The weaker the spring, the more the cone will move for a given current through the voice coil. (Above resonance, the response becomes mass-dominated, because at these higher frequencies the forces required to accelerate the cone become progressively higher. Note that this behaviour is analogous to that of the electrical tuned circuit where the impedance is capacitive below resonance and inductive above. Capacitance and stiffness represent methods of storing potential energy; inductance and mass are reservoirs of kinetic or motional energy). Extended bass response implies large cone excursions at low frequencies, hence low stiffness in the suspension, which is in turn reflected in a low resonant frequency. QED.

Sealed Boxes

We shall begin our survey of loudspeaker enclosures by looking at the hermetically sealed box, represented by the ubiquitous "bookcase" loudspeaker. This enclosure (Fig. 2) is an approximation to the infinite baffle in that it prevents pressure waves from the rear of the cone cancelling those from the front. A typical sealed box design has the loudspeaker mounted off-centre in one wall, and absorbent wadding inside to reduce standing waves set up by reflections from the inner surfaces. However, unless the box is extremely large, it is most definitely not like the true plane baffle of large dimensions that we considered earlier. This is because the movement of the cone results in compression and expansion of the enclosed air, which gets hotter as a result. Thus the amplifier has to deliver extra power to heat the air and, for a given input level, less power is radiated as sound. At low frequencies, an additional aspect of inefficiency also comes into play: this is due to the air loading on the cone, raising the resonant frequency of the loudspeaker by virtue of its stiffness or resistance to compression. We have just seen why this will reduce the radiating efficiency at low frequencies below resonance. (For this reason the suspension of a loudspeaker for use in a sealed enclosure is sometimes very "floppy", so that when in the box a proper value of stiffness is restored by the air loading).

Sealed enclosure design to achieve good bass response is seldom as effective as the advertisements suggest. The story goes that by making the loudspeaker cone able to execute large movements, adequate bass response can be achieved at the expense of the efficiency of converting electrical into acoustic power. To a degree this is true, but the principal limitation (aside from the very real danger of overdriving the loudspeaker with high-power long pedal notes on the organ) is that harmonic distortion occurs owing to the voice coil moving through a not perfectly uniform magnetic field during its excursion. Such distortion results in audibly impure bass for organ work, though the system is perhaps generally acceptable for other types of programme such as tympani transients. Notwithstanding this, though, I have yet to come across a sealed box which is adequate at 30 Hz.

Bass Reflex Enclosures

A reflex enclosure (Fig. 3) is similar to the sealed box except for the presence of a port close to the loudspeaker. The port is usually the aperture of a pipe which extends back inside the box. The air in the box and port has mass and springiness and thus both can resonate at particular frequencies. The loudspeaker unit itself also has its own intrinsic resonant frequency. The trick in bass reflex design is to juggle these resonant frequencies so that, at low frequencies, the pressure waves from the rear of the cone emerge through the port in phase (not in opposition) with those from the front. This is not an easy thing to achieve because of the criticality of the adjustments (Q factors are surprisingly high at these low frequencies for one brought up on a diet of resonance in lossy electrical circuits). This makes it necessary to adjust each cabinet to the characteristics of the drive unit actually used because manufacturers make little attempt to construct loudspeakers to close tolerances. Simply building a reflex cabinet from a published design will not automatically provide good bass no matter how good the original calculations were. There also has to be very effective sound absorption inside the box at frequencies above, say, 70 Hz otherwise the response at higher frequencies will be seriously affected. With all these difficulties, what advantages are conferred by the bass reflex system? Firstly the efficiency is higher than for the sealed box because the energy from the rear of the cone is used rather than being dissipated as heat. It is also possible theoretically to extend the bass response below the resonant frequency of the loudspeaker unit, but this is only practicable with large enclosures. Nevertheless the reflex enclosure does not affect the resonant frequency so adversely as does the sealed box even for a small cabinet

My experience is that bass reflex enclosures are fine in theory but can be disappointing in practice. I spent hours using probe microphones to carefully adjust an 8 cubic foot cabinet containing a 12 inch loudspeaker and obtained a reasonably flat response down to about 40 Hz, but this was a bit disappointing for such a large box. It was useless for organ applications because of the other major shortcoming of the reflex enclosure, which is that the response really does rocket off at the bottom end: 18 dB per octave compared with 12 dB for most other forms of enclosure. Thus my cabinet; fine at 40 Hz, was useless at 30.

Do not assume that the bass reflex enclosure is an old fashioned scheme seldom used nowadays. Commercial designs using the so-called auxiliary bass radiator (ABR) are the reflex enclosure in a new guise. The ABR is placed in the port, being similar to the cone and suspension system of a normal loudspeaker. Its characteristics enable a lower resonant frequency to be obtained from a relatively compact pipe and cabinet. However, small reflex enclosures are more critical to adjust than large ones, and the presence of the ABR makes things one stage more difficult (quite apart from the additional distortion produced by possible nonlinear effects in the ABR, which has cone excursions quite as large as those of the loudspeaker itself). Thus only the most expensive loudspeakers of this type having a frequency response curve supplied can be relied upon to fulfil their brochure promises.

I have indeed invested in a pair of speakers of this pedigree comprising an 8 inch bass unit with a somewhat larger ABR. The major cabinet dimension is about 2’ 6’’ high, yet even from this structure the bass response (from the curve supplied) begins to fall off rapidly at 50 Hz. This is not a criticism of the loudspeaker as such: it only demonstrates that you simply cannot get good bass from objects of this size no matter how expensive they may be. The units are very acceptable for general hi-fi use but completely useless with the lower pedal notes of my electronic organ.

Horns

A horn (Fig. 4) is a flared tube whose diameter at one end corresponds to that of the loudspeaker used as a drive unit, and at the other it is as large as can be comfortably accommodated. In practice the tube might be folded to conserve space. By gradually increasing the width of the horn the relatively stiff cone is matched gradually into the outside world consisting of the tenuous air. This acoustic transformer effect occurs because the progressively widening air column exhibits progressively less inertia.

Thus there is a better impedance match than would be the case for the drive unit radiating directly into the room. This enables greater acoustic powers to be achieved with a given electrical drive level, but only within the frequency region for which the horn has been designed.

The so-called horn cutoff frequency defines the lowest frequency for which "horn action" will occur. It really disguises the fact that the mouth of the horn is a radiating aperture just like a normal loudspeaker mounted in a baffle, and we have already seen how bass response depends on the size of the aperture. A horn cannot be flared too rapidly or it ceases to act like a horn. The flare usually follows an exponential or similar curve and this limits the horn aperture of many designs to an area roughly equivalent to a 15 inch cone if the horn is to be constrained to a length suitable to a domestic room. Consequently the bass reproduction of such a horn will be no better than a single directly radiating 15 inch loudspeaker, and worse than two 12 inch units driven in parallel.

The issues of efficiency and drive level for horns need to be explored a little further. With horn sizes that would fit in an average domestic room it is unlikely that the horn cutoff frequency will be as low as we would wish – 30 Hz is our target figure. Therefore at these low frequencies the horn would not in fact be acting like a horn in an acoustic sense, and the acoustic transformer effect will be degraded or non-existent. This could mean that a small driver unit at the throat of a horn might have to move a volume of air equivalent to that which would be moved by a larger loudspeaker of equivalent area to the mouth. Consequently the little driver may have to work very hard in terms of cone excursion and therefore it may generate more distortion than the larger direct radiating loudspeaker. It must be remembered that, for horns, the virtues of increased efficiency and better bass response only refer to the size of loudspeaker used as a driving unit above the cutoff frequency: if a horn is driven with a 5 inch unit and has an aperture equivalent to a 15 inch unit, then certainly the efficiency and bass response of the small speaker is enhanced to approximate that of the larger one. However below cutoff the combination of such a horn and driver is no more effective than a single 15 inch direct radiating loudspeaker in a suitable cabinet would be, and may well be worse in terms of harmonic distortion (I will not at this juncture define what might constitute a "suitable cabinet", I only would point out that it is likely to be less complex and less domestically objectionable than a horn).

The conclusion is that horns are only justified where advantage can be taken of the cone area magnification effect to achieve an equivalent area to an otherwise unattainable loudspeaker or loudspeaker array. In these circumstances the horn will be truly enormous, suitable only for concert halls and the like. The lingering penchant for horns in a domestic environment, it seems to me, is a hangover from the days when it was not easy to obtain large, good quality direct radiating loudspeaker units and it was certainly not possible to drive them with more than a few watts of power.

Other Designs

There are several other types of loudspeaker enclosure that might be discussed but none are particularly notable for their extremely low frequency performance. One enclosure allows the front of the loudspeaker to radiate direct into the room whilst the energy from the rear is dissipated in a long pipe, usually folded to conserve space.

Acoustically absorbent material is placed in the pipe to prevent sound emerging from the end and interfering with that radiated from the front of the cone. Unfortunately at low frequencies the effect of such absorbers is negligible, thus there will be frequencies at which phase reinforcements or cancellations occur in the listening room. Also the mass of air in the box and pipe, even though the pipe is open, raises the resonant frequency of the system. For low frequency reproduction this design therefore has little to recommend it

The Effect of the Listening Room

The excitation of standing waves caused by reflections in the listening room is a well known phenomenon. In a rectangular room a standing wave will be set up between two parallel walls at frequencies corresponding to multiples of half a wavelength. This is a form of resonance and a large number of resonant frequencies can occur in practice. Some of these correspond to transits around the room involving various surfaces at various angles of incidence. Thus not only is the number of resonances potentially very large, but even relatively small rooms can exhibit resonances at low frequencies where a "round-the-room" reflection path is involved. In a medium sized room with one dimension around 18 feet (half a wavelength at 30 Hz) a pronounced resonance at this frequency is usually subjectively obvious. The position of the loudspeaker in the room has some influence on the number of standing wave resonances that are induced, and a corner position will excite more of them than one close to the centre.

Bearing in mind the difficulty of acoustically characterising a real room, as distinct from the closed rectangular enclosure beloved of the theoretician, I doubt whether there is much to be gained from pursuing the standing wave situation much further. Many real rooms may indeed be substantially rectangular, but the presence of windows (open or closed depending on the season), one or more doors (ditto), curtains, carpets and furniture together with the largely unknown reflectivity characteristics of the surfaces all suggest to me that we have a situation not very amenable to analysis. Consequently I proceed on the premise that we have to put up with the room we have, and in any case the position of a large low frequency loudspeaker is usually dictated by other than scientific considerations. Thus before moving on I shall merely emphasise that room resonances **will** occur, and this means that subjective low frequency effects will vary with listening position even in a small room. If the room is much smaller than half the wavelength (say 18 feet at 30 Hz), the standing waves will tend to cancel out any energy. If you tried to listen to 32 foot notes in such a room it is likely you would hear nothing at all, except possibly in quirky and extremely critical positions where complete cancellation had not occurred.

What is the best system?

Having read this far you must have tired of my dismissive criticisms of all the loud-speaker designs discussed. A legitimate question at this point, therefore, is where do we go from here? Until relatively recently I was still asking myself this question, having spent years and a lot of money in either purchasing, constructing or borrowing examples of all the enclosure designs previously referred to. In spite of various claims none of them proved satisfactory in being able to imitate that subtle sensation perceived in a church when the soft pedal stops are in use. This situation changed dramatically when I moved into a house that allowed my electronic organ to be installed in a room having its own roof space accessed by a trap door in the ceiling. It occurred to me that here was a near-perfect infinite baffle. The ceiling of the room, although made only of plasterboard, was braced at regular intervals by massive baulks of timber (the ceiling joists). Would that our normal loudspeaker cabinets were built like this! The volume of the roof space and the fact that it was (acoustically) open to the outside world meant that there would be no compressive loading on the rear of the loudspeaker cone as would have been the case with a sealed box. The fortuitous presence of the access hatch actually in the room itself provided the final temptation to mount a pair of 12 inch speakers in the trap door. (Remember that a pair of speakers is four times as effective as one). The arrangement is sketched in Fig. 5. The loudspeaker units had a free air resonance of about 35 Hz, and were mounted along a diagonal of the trap door which was about an inch thick.

The results were, to put it mildly, certainly without equal to anything I had previously experienced. Winding a sine wave oscillator slowly up and down between 20 and 50 Hz produced the "feel" rather than the sounds of these low frequencies and at surprisingly low powers. Only 5 watts of electrical power produced results that rattled doors, windows and radiators even in distant rooms. The lowest frequencies seemed to permeate the entire house. On the electronic organ itself I had to attenuate the outputs of the pedal stops over the lowest octave to prevent quite deafening results. On the softest 16 foot flute stops I was able to obtain the quiet, breathing bass that had eluded me for so long. In fact, one can hear it rather too well in the garden. To back up all this qualitative stuff I did actually measure the low-frequency response with some care (Fig. 6) which tended to confirm my subjective impressions.

> The curve in Figure 6 was fitted to data obtained from capacitor microphones at various places in the room, and the rms error indicated is about +/- 2 dB. The most consistent results were obtained for measurements in the centre of the room, with more scatter characteristic of the corner positions. This was not unexpected from the predicted standing wave effects in a rectangular room. The response is substantially flat down to 30 Hz and it falls off at 18 dB/8ve below that frequency. Compare this with the response of my "hi-fi" speakers (Fig 1b) which even in the 1970's cost several hundred pounds! These are at least 10 dB down at 30 Hz.

Somewhat later I was leafing through Beranek's well known book "Acoustics". I was struck by the fact that he wades heroically through the detailed theory of reflex cabinets, horns and the like and, at the end of it all in a hidden little paragraph, mildly suggests that for good bass one should try to create a true infinite baffle preferably using multiple loudspeaker units. The method he suggests is to mount the speakers in the wall of "a closet filled with clothing". Obviously the great Beranek must have had a similar revelation when he tried it!

In spite of the jokey images, I am quite seriously suggesting more widespread use of the roof space as a means to achieve an infinite baffle, though obviously I do not claim to have invented the idea. The practical difficulties are not trivial, particularly if a trap door has to be constructed for the purpose. It is also difficult to feed power to loudspeakers in this position and perhaps you would be willing to forego a power point in the room so that the heavy-gauge cable already in the walls can be used. Finally, and again seriously, you must protect the speaker cones (if of paper) from the attentions of mice. I surrounded the units with tubes of corrugated cardboard topped by plastic garden mesh.

The ceiling loudspeaker system can of course be used for wide range reproduction from a suitable loudspeaker, but if more than one unit is used it is advisable to restrict operation to low frequencies only. Multiple loudspeakers will produce interference peaks and nulls in the response at medium frequencies and above, and in my case no frequencies above about 170 Hz are fed to the ceiling loudspeakers.

Another method of creating an infinite baffle is to mount the speakers in a wall separating one room from another. Commonly such walls will be studwork partitions, thereby reducing the labour involved, and whilst acoustically less attractive than a solid wall of brick or blocks they would probably be at least as effective as the ceiling mounted system. Alternatively doors or cupboards could be used, a la Beranek. Whilst at first sight these techniques may sound crazy, it has been the object of this article to show that nature demands extreme methods for the reproduction of extreme bass. The methods described seem no more outrageous than (for example) the folded concrete horns employed by other of the cognoscenti!

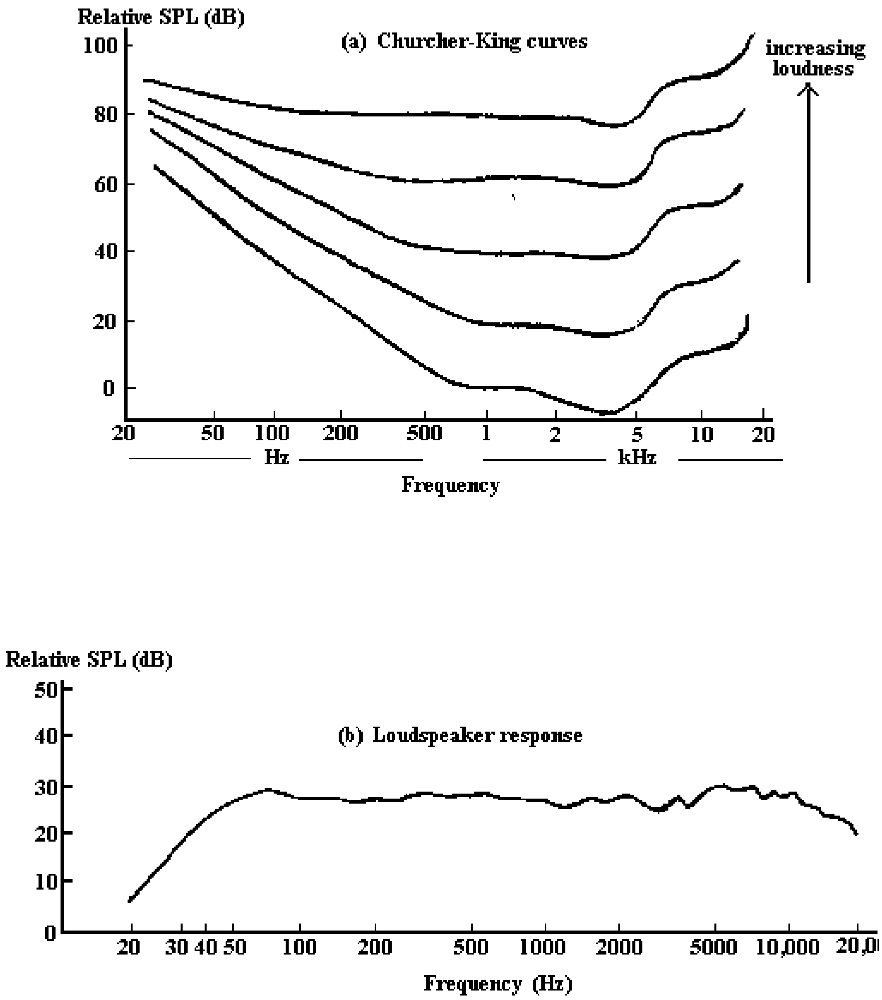


Figure 1. Problems of radiating low frequencies

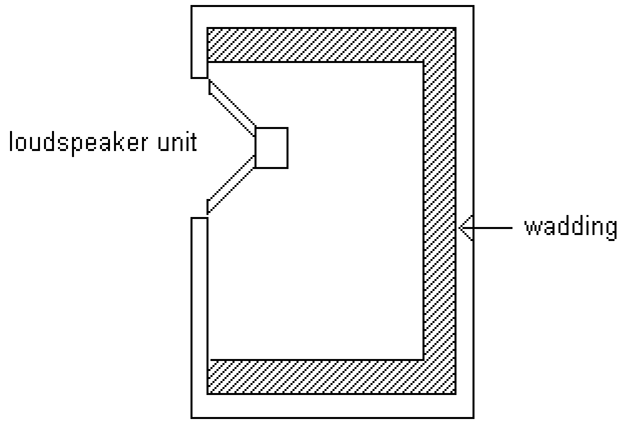


Figure 2. Sealed box enclosure

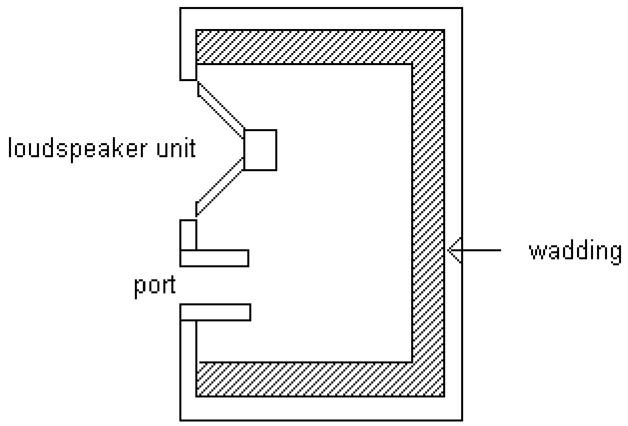


Figure 3. Bass reflex enclosure

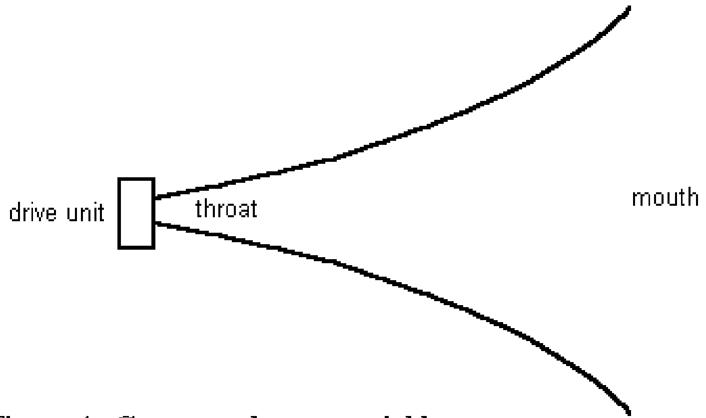


Figure 4. Conceptual exponential horn

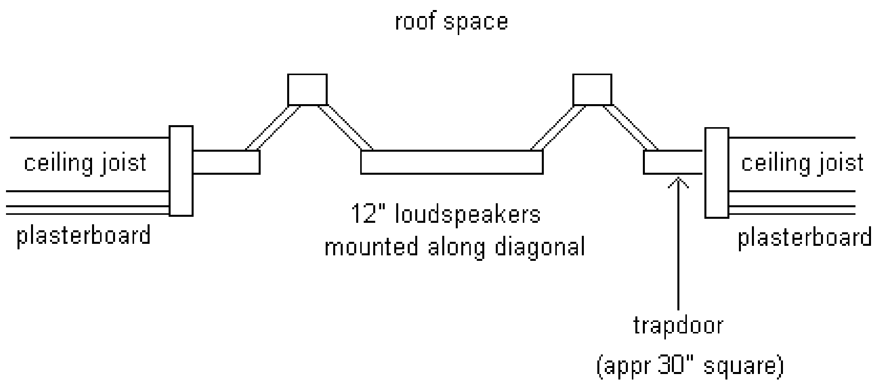


Figure 5. Ceiling mounted loudspeakers

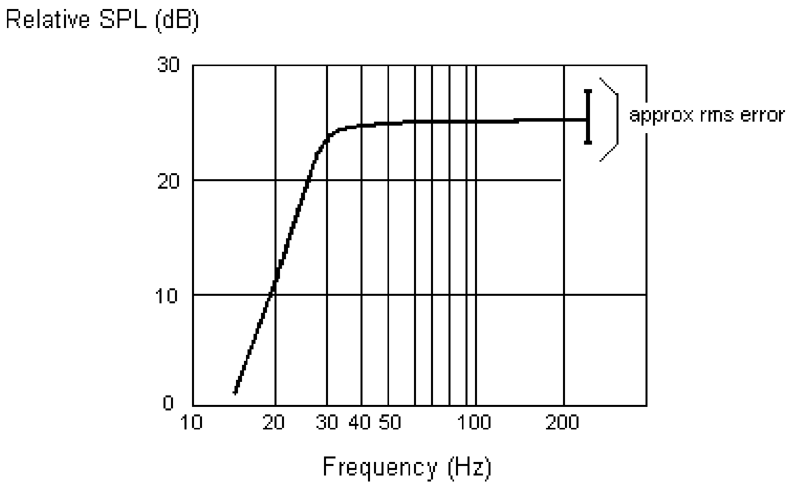


Figure 6. Measured response of ceiling loudspeakers

APPENDIX 2

Phase Shift Vibrato and Chorus

by Colin E Pykett

(This article first appeared in EOM 178, July 2001)

There still seems to be a substantial market for audio effects circuits, judging by the frequency with which they appear in hobby magazines and books, and complete units are still widely advertised. This article describes another look at an old idea, that of obtaining vibrato and chorus effects using a phase shift circuit. An advantage of phase shift circuits is their simplicity – only a few ordinary transistors are required rather than special purpose devices such as bucket brigade delay lines or digital signal processing in a computer.

Because of this simplicity phase shift vibrato was popular in the early days of electronic music, and it produced effects we do not often hear today as will be described later. Although a constant phase shift has no audible effect on a signal, if the phase shift is constantly varied the instantaneous frequency of the signal also varies. Therefore a cyclical phase shift circuit produces a cyclical frequency modulation effect, which causes the subjective perception of vibrato. (Phase modulation and frequency modulation are similar, but not exactly the same thing. Being rigorous about it, frequency is proportional to the time derivative of phase, but we will forget about this for the purposes of this article).

Early Hammond organs used phase shifters to obtain vibrato and chorus effects, but these were elaborate transmission line systems using large numbers of inductors and capacitors and a mechanical scanner. An effect similar to that obtained from today's electronic delay technology was obtained, but things had to be done that way in those days of course. An early form of purely electronic phase shift vibrato, using valves, was described by Alan Douglas in *The Electronic Musical Instrument Manual*, 4th edition, pages 256 and 265. Recently I decided to design a transistor version of it, starting from first principles, and it is this circuit which will now be described.

Let us begin with the classical phase shift network, the analysis of which still forms part of the staple diet of unfortunate electrical engineering and physics students. While it is necessary to understand what the network does, we do not need the associated maths here. Referring to Figure 1, the input signal is applied as shown and an output obtained by tapping the two arms of the network. The network has an extremely useful property. If R is varied, the amplitude of the output does not change but its phase relative to the input does, over a range approaching 180 degrees. The amount of phase shift also depends on frequency, decreasing as the frequency increases. Thus if we apply a signal containing many frequencies, such as the output from an electronic musical instrument, the phase shift will be greatest at the lowest frequencies but the amplitude of each frequency will not be affected.

Therefore for any constant value of R, no audible change will occur in the output. This is because each frequency component has not been altered in relative amplitude, only phase. Because the ear is not sensitive to unvarying phase, no audible change will result. (Even so, because the phases of the various frequency components have been changed, the waveform of the output will not look like the input). But if R is instead made to change cyclically with time, we will hear a vibrato effect, with the depth of vibrato increasing towards the lower frequencies.

There is unfortunately a snag with the phase shift network, which is that the variation of phase shift with the value of R is highly non-linear. The phase varies rapidly for small values of R but slows down as R becomes larger. Without compensating for this, the vibrato effect would be correspondingly non-linear and it would sound unpleasantly thumpy. Another practical problem is how to control the value of R electronically in the first place. All sorts of methods have been tried by others, such as light dependent resistors and voltage dependent resistor elements. But if one uses a junction field effect transistor (JFET), both of these problems are solved simultaneously.

A JFET can be used as a simple voltage dependent resistor, whose resistance depends on the negative bias applied between gate and source. This variation, too, is highly non-linear, in that the variation of resistance with voltage is slow for small values of bias but becomes much faster as the bias reaches a few volts. (For a more detailed description of JFET characteristics see my article in EOM 165, February 1998). By proper design the nonlinearities of the phase shift network and the FET can be made to approximately cancel, giving a voltage-controlled phase shifter which is fairly linear over quite a wide range. Typically 120 down to 20 degrees of shift at 300 Hz can be obtained for a control voltage variation of 0 - 2 volts negative.

A circuit using these techniques is shown in Figure 2. The cyclical variation is derived from an RC oscillator, and by suitable choice of components it will deliver either vibrato (modulation frequencies of a few Hz) or chorus (modulation frequencies less than one Hz). Typical component values for the oscillator are as follows:

FUNCTION	**C**	**R**	**R1**
Vibrato	4u7	1k0	470k
Chorus	22u	10k	220k

Note that two stages of phase shift are used. However only a single stage might be adequate, depending on your subjective reactions. I suggest building it first with just one stage to assess its effect.

The following design features are important:

1. The DC level preset should be adjusted so that the output from the oscillator, as seen by the gates of the FETs, does not go above zero volts when maximum modulation depth is used.

2. Vibrato rate can be adjusted by making one of the R’s variable.

3. For best results the minimum resistance of the FETs should be chosen to be as low as possible. This parameter is not well controlled during manufacture so a small batch

of 5 or so should be tested. The gate and source leads should be temporarily joined together, and a resistance meter applied between drain and source. The positive terminal of the meter should be applied to the drain and the negative to the source. (Note that in the simple form of meter which uses an ammeter in series with a battery, the positive pole will usually be the black lead). FETs with resistances of about 300 – 500 ohms should be sought. The reason for being careful about polarity is not because the FET would otherwise be damaged (which it won't), but because different readings are obtained with the two polarities.

4. The input signal should come from a very low impedance source otherwise the phase shift network will be unbalanced and will not work properly. Even a source impedance of 100 ohms is too much, so the safest method is to buffer the input signal with an operational amplifier before applying it to the circuit.

The subjective effect of this vibrato is different to that which we usually encounter today. The frequency modulation from delay lines or from computer organs is usually mathematically perfect, being accurately sinusoidal and operative over a wide frequency range. This means that if one wants to get a heavy wobble on tones with little harmonic development, such as flutes or tibias, one often then finds the modulation is far too deep with reeds and strings. By contrast, this circuit produces a depth of modulation which increases towards the lower frequencies. It can therefore produce a really heavy wobble on flutes if necessary without it becoming objectionable on other combinations. It therefore has an attractive retro sound of the sort produced by some old drawbar organs, and it approximates quite well to the tremulants found on some pipe organs. It might not provide quite the amount of "giggle" needed for Wurlitzer tibias or fairground organ flue work in their upper registers, although this is a matter of taste.

The fact that modulation depth decreases with increasing frequency also helps when the circuit is used as a chorus generator. True frequency modulation, as produced by delay line technology for example, tends to produce far too much phase shift at high frequencies when adjusted for satisfactory operation at low to moderate frequencies. The resulting effect can make the higher pitches from mixtures, etc, drift distinctly in and out of tune to an unpleasant degree. This circuit can be set up to produce a sense of "movement" in the sound which is just enough to unlock the phases of different signal sources, without actually detuning them at high frequencies.

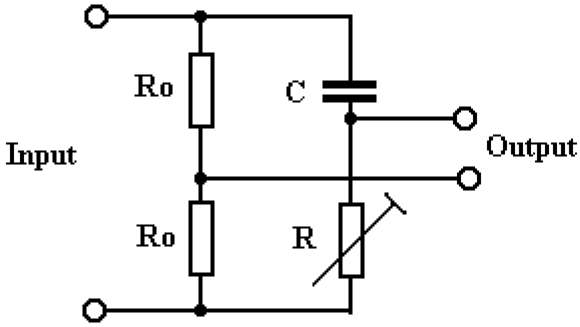


Figure 1. Phase shift circuit

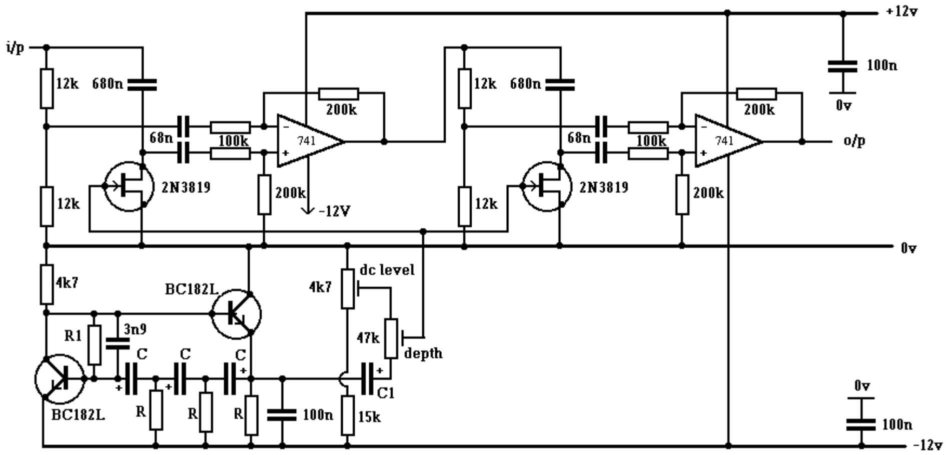


Figure 2. Complete phase shifter circuit

APPENDIX 3

A Multichannel Chorus Processor

by Colin E Pykett

(This material was first produced in 1985. It uses some components which are now difficult or impossible to obtain, in particular the TDA 1022 bucket brigade device)

INTRODUCTI ON

This information package describes a multichannel processor which slowly shifts the phases of a number of audio frequency signals back and forth with respect to each other in an unsynchronised fashion. Thus signals emerging from different channels are heard to “beat” even if they originated from a single source. When applied to the outputs of various tone filters in an electronic organ the processor produces an effect similar to that obtained in a pipe organ where the various ranks of pipes are seldom exactly in tune. The effect is also very similar to that from a multi-rank electronic organ. The processor is most useful for “unlocking” the phases of stops which are derived from the same generator rank. For example, in an organ with a single rank of generators, the 8 and 4 foot diapasons could be processed by different chorus channels to simulate the presence of two ranks. With the multiple channels available (10 in this design) this could be done with many other combinations of stops as well. Certainly with a one rank organ this processor will add a liveliness to the sound that is well worth the investment in building it. With an organ already having multiple generator ranks the processor can be used to unlock the phases of, say, all the 8 foot stops which will commonly be derived from one rank. Only on an organ having a separate rank for each and every stop will this processor be redundant.

What is involved in building it? Basically there are 8 chorus channels and 2 “straight through” channels, making a total of 10 signal paths. The main cost and effort lies in the chorus channels themselves, which at the time of writing each require about £6 worth of components depending on where you obtain them. Thus the total component cost of the complete system will be in the region of £60. The prototype system ended up on 12 5” by 4” Veroboards but this excludes the swell pedal, tremulant and reverb circuitry which is not part of the chorus system and which is not described here.

OUTLINE DESCRIPTION (refer to Fig 1)

The system begins with a patchboard which will accept as many audio inputs as there are stops on the organ. It can be expanded as necessary but the prototype had a capability of 40 stops and it was designed with a 2 manual and pedal instrument in mind. The patchboard is used to select which stop is routed through which chorus channel, a facility which enables different configurations to be tried. For example, if you have only a single generator rank, then you would probably wish to have 8, 4 and 2 foot diapasons (say) going through different chorus channels since this is a common stop combination with which the unlocking effect of the processor would be exercised. If, on the other hand, you have three separate generator ranks for the 8, 4 and 2 foot stops then the processor is not needed to unlock the phases of stops drawn from these ranks. You would then tend to use it to unlock all the 8 foot stops, all the 4 foot stops, etc, or at least common combinations of them.

The ten patchboard outputs are next fed into the eight chorus channels and the two direct or straight-through channels. A total of five channels (4 chorus plus 1 direct) is reserved for the swell department and the other five for the great and pedal. This subdivision is somewhat arbitrary and could be varied to suit, say, a three manual scheme. In principle there is no need to make such a separation at all at this point in the system, but it is useful to separate out at least the swell department since it will have swell pedal and tremulant circuitry applied later whereas the great and pedal will not, at least in a classical organ.

The direct channels are used partly for economy and partly for technical reasons. On economic grounds it is not necessary to have a chorus channel for every signal path: the combination of two signals routed through one direct channel and through one chorus channel is more or less as effective as if both channels used chorus. On technical grounds it is useful to have a direct path for loud stops, particularly those on the pedal, which might otherwise overload the bucket brigade device in a chorus channel; these devices have limited dynamic range and a most unpleasant overload characteristic.

The channels are labelled 0 to 4 for the swell and 100 to 104 for the great and pedal. Channels 0 and 100 are the direct ones, all others use chorus.

After chorus has been impressed on the signals in the eight channels 1 to 4 and 101 to 104 the signals pass through clock frequency rejection filters to prevent spurious low level whistles that might otherwise occur owing to interaction between the supersonic clock frequencies used in the chorus channels. The direct channels (0 and 100) do not require such filtering. The ten signals are then routed to a second patchboard which enables them to be individually allocated to main amplifier channels. In this design four such output channels are used, two for the swell and two for the great and pedal.

It is highly desirable to have as many sound output channels as possible. The subjective effect of chorus is better if the various outputs are allowed to mix acoustically, by being fed to different loudspeakers, than if they are mixed electrically and then fed into a single speaker. Power amplifiers and loudspeakers are expensive, of course, and the provision of four output channels is a compromise between the ideal (which would be one per chorus channel) and economy. The patchboard allows the chorus channels to be fed to different loudspeakers in a flexible manner.

Following the patchboard the two “swell” output channels pass through swell controls and tremulant circuits (not described). The “great” and “pedal” channels do not, though there is no reason why they should not if you so wish. The four output channels then pass into a preamplifier board which also contains provision for driving a reverb unit (also not described here). After the preamps the signals are available for power amplification.

DETAILED DESCRIPTION

Input Patchboard

This is illustrated in Fig 2. Up to 40 stops, tone filter outputs or other signal sources are fed into the patchboard through a set of 22k resistors. The output impedance of each signal source should be low compared to 22k and be symmetrical about zero volts (i.e. no DC component). Using an op amp in each source output is the neatest way to meet these requirements. The 22k resistors feed a 400 way cross point matrix such that by making the appropriate connection a stop can be fed into any one of the ten output channels. Each stop must only be fed to one channel - this is important. The output channels have a simple summing amplifier to add the contributions from the various stops that have been assigned to a particular channel, and to present a low impedance to the chorus channels which follow.

The construction of the patching matrix requires some discussion. Without doubt, the most elegant solution would be to use commercial crossbar matrix blocks in which the assignment of stops to output channels would be done by inserting pins in the appropriate holes. One very compact 10 x 10 block is the RS Components 468-024, of which four would be needed to accommodate 40 stop inputs. However the set of four, excluding pins, would currently cost about £43 (incl VAT) at trade prices. A much cheaper alternative is to assemble an array of Vero pins on Veroboard or similar, and solder a flying lead from each 22k resistor to the pin representing the desired output channel. It is possible to short circuit some of the effort by noting that stops on the swell only need to go to channels 0 to 4, and on the great and pedal to channels 100 to 104. This cuts down the number of pins needed in the matrix. Using this approach I assembled the complete patchboard, including buffer amplifiers, on a piece of Veroboard about 5” x 4”.

Electrically, the only critical layout feature concerns the ten output lines of the matrix, those running vertically in Fig 2. Note that these are the summing nodes of the respective buffer amplifiers. The use of the term “virtual earth” leads to a common misconception that these are low impedance or even earthed points. In fact these points in the circuit are associated with very high gain at an effective input impedance of up to 22k, depending on how many inputs are connected to the matrix. Thus the amplifiers should be located as close as possible to the patching matrix to prevent possible pickup of hum and spurious signals; if this is not possible then screened lead should be used to make the connections from the matrix outputs to the summing amplifiers.

Chorus Channels

There are eight identical chorus channels, having the functional block diagram in Fig 3. The input signal (channels 1-4 and 101-104) from the input patchboard is low pass filtered at 15 kHz to prevent aliasing in the bucket brigade (BB) device that follows. (If the input frequency is too high relative to the clock frequency used in the BB, spurious or “aliased” tones will appear in the output). In the BB device the clock frequency is modulated by a very low frequency oscillator. The modulation is approximately linear period, that is, the period of the VCO or clock oscillator is proportional to the instantaneous voltage from the VLF modulating source. This produces linear frequency modulation in the signal at the output of the BB chip. Since the VLF oscillator is working at very low frequencies, around 8 seconds period (1/8

Hz), the frequency modulation is too small to detect on a single sound source but it produces pleasant beating effects between the signals emerging from the various chorus channels. The amplitude and frequency of the modulation can be adjusted to taste. Finally the signal passes through a second low pass filter to remove much of the clock frequency components. Additional filtering, on the belt-and-braces principle, is provided later in the system.

The circuit of the signal channel is at Fig 4, and generally follows the guidelines in the data sheet for the TDA 1022 BB device. Thus both of the low pass filters are Rauch second-order circuits with capacitors returned to the positive line which is connected to the substrate of the chip. Separate potential dividers are used for setting the DC levels at pins 5 and 13 rather than a single one as in the data sheet. This makes final setting up easier.

The clock oscillator circuit is at Fig 5. The VLF modulation source is based on a design by Ryder and published in ‘Wireless World’ in March 1979 (page 53). Presets VR1 and VR2 control the frequency and amplitude respectively of the sine wave passed on to the VCO or clock oscillator. Ideally the feedback capacitor across the second 741 should be non-electrolytic (2.2 microfarad) and these were used in the prototype. However the solution of back-to-back tantalum electrolytics as in the diagram should work in practice and will be cheaper.

The VCO is a variant of a circuit described in the data sheets and it uses a single 4093 CMOS device (Schmitt trigger) as a period modulated oscillator. Because of variation in the trigger levels between different 4093 specimens it is necessary to select the timing resistors (R3) to bring the clock frequency into the correct range. Once having done this the circuit remains very stable. The setting up procedure, which includes defining the clock frequency, is described later.

Construction of the circuit is generally non-critical and the whole chorus channel can be comfortably accommodated on a 5” x 4” Veroboard. Vero pins should be used for R1, R2, R3 (2 resistors), R4 and C1 as these need to be adjusted on test. Additionally it is useful to use a pair of pins to enable point A in the oscillator to be temporarily disconnected whilst setting up. Holders should be used for the 4093 (14 pins) and the TDA 1022 (16 pins).

Setting up procedure

After building the circuit do not insert the TDA 1022, and disconnect link A. Check for obvious faults such as shorted power lines and poor soldered joints, then:

1. Use 82k resistors for R3 and check that antiphase square waves appear at pins 10 and 11 of the 4093. Select R3 until the frequency is about 40 kHz (25 microseconds period).

2. Switch off and temporarily connect a 1k pot between V_A and ground, with the slider connected to point A (the two diodes). Set the slider to ground and switch on. Slowly increase the voltage from zero and observe that the output frequency remains constant at the previously set value (approx 40 kHz) until a certain threshold voltage is reached. Note this voltage (V1). Continue increasing the voltage until the oscillator latches up and stops oscillating - note

this voltage also (V2). Switch off, remove the pot and reconnect link A.

3. Calculate a value for R4 such that the voltage at pin 3 of the corresponding 741 is midway between V1 and V2. The following equation will give an approximate value:

$$R4 = \frac{10V2 + 10V1}{24 - V2 - V1} \text{ (in kohms)}$$

If necessary use two resistors in series.

4. With this value of R4 set VR1 and VR2 to maximum resistance and switch on. Check first that the polarity of Cl is correct; some values of R4 may necessitate a change-round. Then adjust VR2 until the clock frequency is varying between its lower value of 40 kHz and just below latch up. Difficulty in achieving this condition may indicate that R4 should be altered slightly. An accurate choice of R4 will result in the two limits above (40 kHz and latch up) being reached simultaneously at the same setting of VR2. VR1 can then be used to adjust the modulating frequency as required (typically between 5 and 10 seconds period). VR2 will control the modulation amplitude but the setting just derived will suffice until the whole system is working and you have developed more familiarity with it.

5. Switch off, insert the TDA 1022, and apply a 250 Hz sine wave at about 1 volt p-p to the signal input. Temporarily connect a 100k pot set to mid-range in place of R1 and R2, and switch on. Adjust the pot and the input signal amplitude until the output signal clips symmetrically - this should occur at about 3 V p-p output amplitude which corresponds to an input amplitude of about 1 ½ V p-p. Measure the resistance of the pot and replace it with a fixed combination (R1 and R2) to suit.

The circuit is now set up. The procedure is somewhat involved, but this is partly a result of the simplicity of the circuit which relies on parameters of the integrated circuits for its operation. Monitoring both the input and output signals simultaneously on a 2 beam scope will show the slowly varying relative phase shift, which should reach a maximum of about 1 cycle (360 degrees) at an input frequency of 250 Hz (corresponds to middle C). Play around for a while with VR1 and VR2 to get the feel of the circuit. Excessively high modulation amplitudes (defined by VR2) will drive the oscillator into latch up and output from the TDA 1022 will cease. However the oscillator recovers as the modulation voltage swings down again, though obviously the circuit cannot be used in this mode. Observe that the phase shift is higher for higher input frequencies; this effect produces an exciting scintillation with upperwork on a classical organ. It will be as well to check that there is a net gain of about two through the signal channel, and that cut-off occurs rapidly at 15 kHz. Otherwise the filter circuits are not working.

Clock Rejection Filters

A second order clock frequency rejection filter is included in each chorus channel. Additional filtering is provided here to ensure that no vestiges of the supersonic clock frequencies can reach the mixer stages in the output patchboard. Any residual non-linearity here would cause audible whistles as the eight unsynchronised clocks swing in and out of phase. Thus proper filtering is essential.

The filters are second order Sallen and Key sections with a 15 kHz breakpoint. A passive lag built into the output patchboard (described later) makes the filter third order. Combined with the second order Rauch filters in the chorus channels themselves there is an overall fifth order (30 dB/octave) clock rejection filter between the outputs of the bucket brigade devices and the following stages.

The circuit is at Fig 6. All eight filters can be accommodated on one Veroboard 5” x 4” as used for the previous circuits.

Output Patchboard

The output patchboard is similar to but simpler than the input patchboard already described. Its function is to allocate each of the 10 channels going into it to one of the 4 output channels, via a 10 x 4 matrix. The circuit is at Fig 7. The 2n2 capacitors used in the inputs of channels 1 - 4 and 101-104 form the third pole of the clock rejection filters. Also the different resistor values in the input lines equalise the gains between the two direct channels and the eight chorus channels (which have a gain of about 2).

In the prototype two of the output channels (1 and 2) were reserved for the swell stops only, and these were routed through dual tremulant and swell pedal circuits. The other two channels (3 and 4), for the great and pedal, go direct to the output preamplifiers. The swell and tremulant circuitry is not described here, but preferably it should have a net gain of about unity to remain consistent with the unmodified channels 3 and 4.

Again, the patchboard can be built on a single 5” x 4” Veroboard as used for previous circuits.

Output Preamplifiers

The output preamps (Fig 8) are straightforward and include provision for driving a reverb unit. When the gain presets of the output channels are at maximum the through gain of the entire chorus processor is about 2 (6 dB). Through gain is the gain between a signal entering the input patchboard and emerging from an output preamp. The gain capability of the reverb driver is somewhat higher, and the signal coming back from the reverb unit is controlled by a further preset with a maximum gain of unity. These provisions enable a wide range of reverb systems to be accommodated. Note that the reverb signal is a composite of all four channels, and this in turn is fed into all four loudspeakers at once. Thus the output of channel N (N = 1,2,3 or 4) is direct signal N plus reverb signal from 1+2+3+4. You may wish to vary this, but I have found it most effective.

The outputs of the preamps are symmetrical about 0 volts, thus no coupling capacitors into the power amplifiers are required.

Power Supplies

The unit as described requires +12 VDC at 160 mA and -12 VDC at 60 mA. Both supplies must be stabilised. If the supplies are derived from a power unit already feeding existing circuits the decoupling arrangement in Fig 9 should be used.

CONSTRUCTION

With careful layout all of the circuits described can be fitted onto a number of standard Veroboards about 5” x 4” (Vero order code 200-21072D) having 36 tracks running parallel to the longer dimension. The number of such boards required is:

Input Patchboard	1
Chorus channels	8
Clock rejection filters	1
Output patchboard	1
Output preamps	1
TOTAL	12

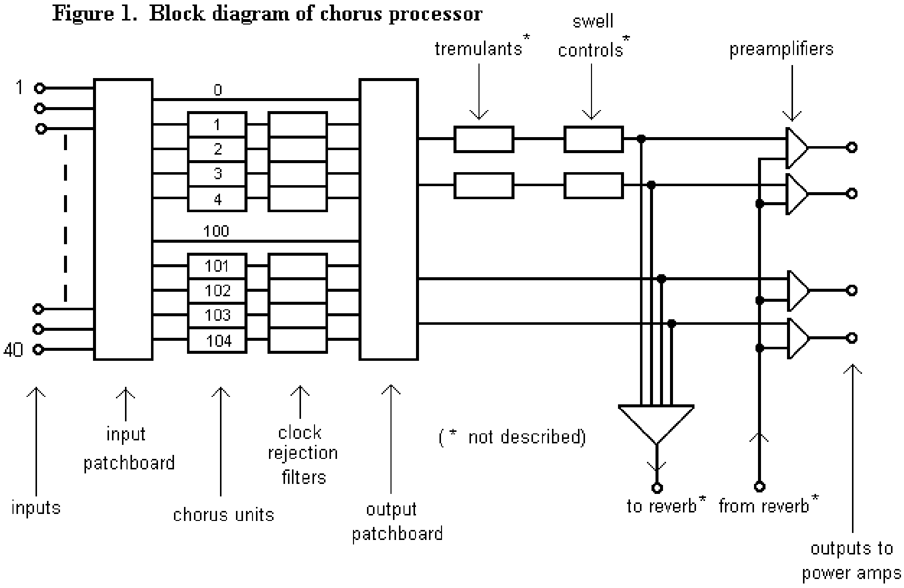
(This does not include swell pedal, tremulant and reverb circuits)

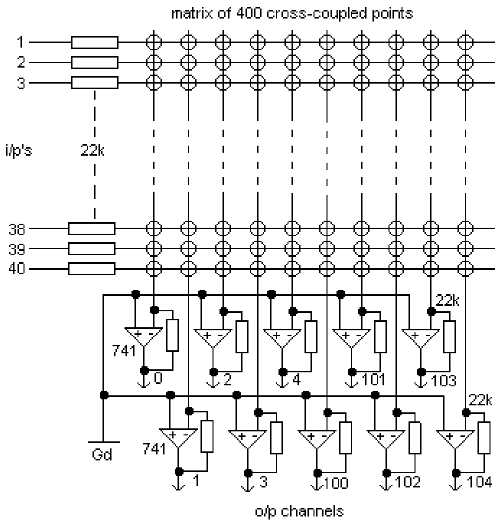
Such a modular construction was used in the prototype. The cards were housed in a home - made frame built from aluminium angle, and interconnections were made using the RS Components PCB interconnector system.

COMMISSIONING

The only parts of the system needing special setting up are the chorus channels, and this has already been fully described. The only other adjustments are the preamp gains for driving the main amplifiers and the reverb unit. Input signals from the tone sources (i.e. at the input patchboard) should not be too small otherwise an inferior signal to noise ratio will result. Input signals around 1 volt p-p are recommended. As experience is gained with the system you may wish to vary the chorus characteristics using VR1 and VR2 in Fig 5.

Figure 1. Block diagram of chorus processor





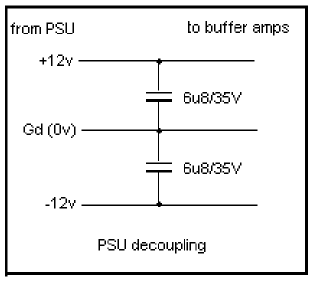
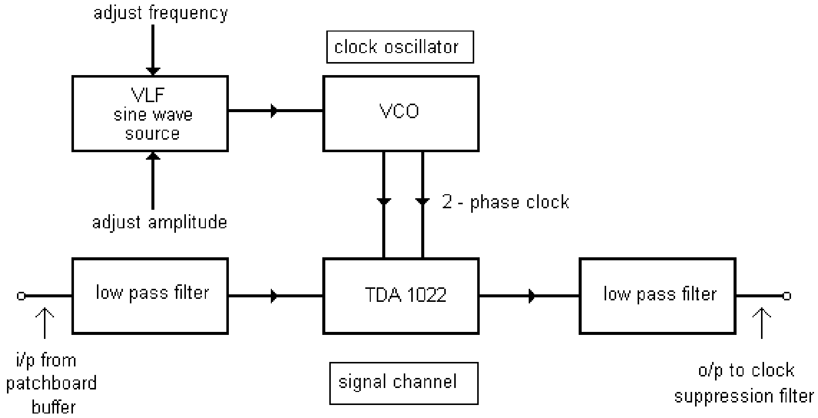


Figure 2. Input patchboard

Figure 3. Block diagram of one chorus channel



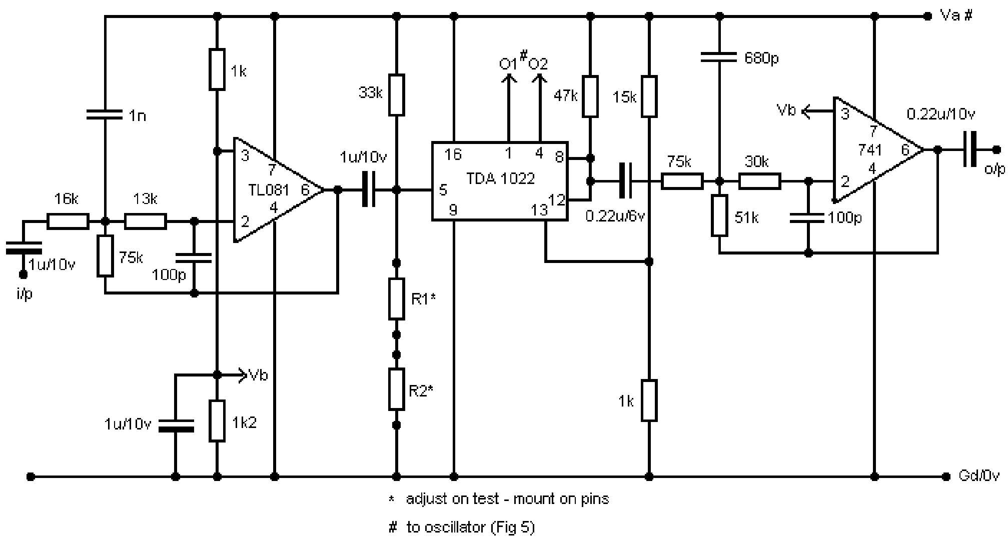
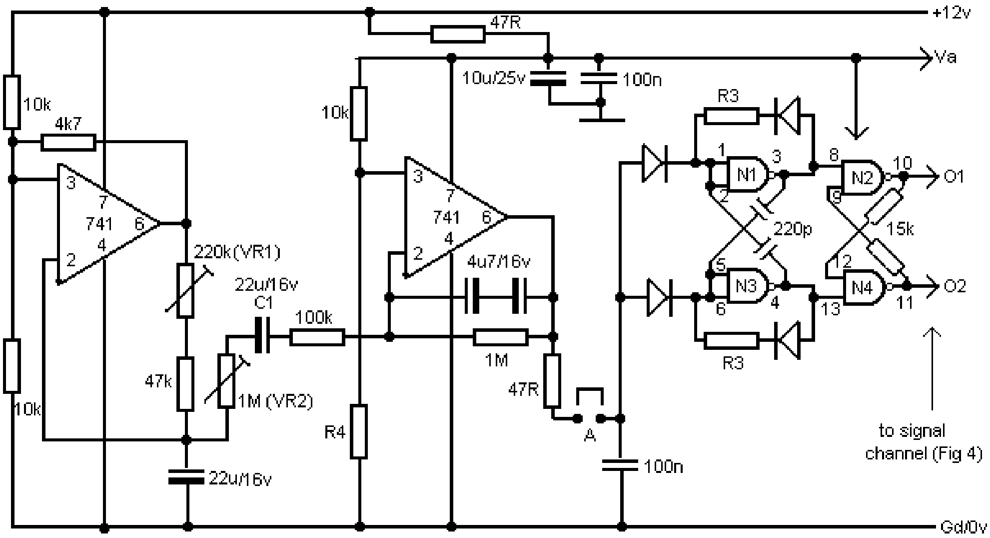


Figure 4. Chorus - signal channel

Figure 5. Chorus - clock oscillator (VCO)



NOTES:

1. N1 - N4 = 4093 (operates between Va & 0v)
2. Diodes = 1N4148
3. R3 (2 resistors), R4 & C1 : adjust on test (mount on pins)
4. VR1 : adjust modulating frequency
5. VR2 : adjust modulating amplitude
6. Point A : link used during set up (use pins)

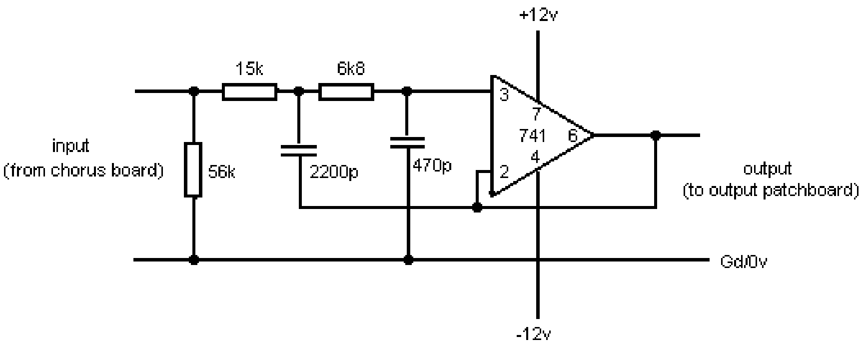


Figure 6. Clock rejection filter (8 required)

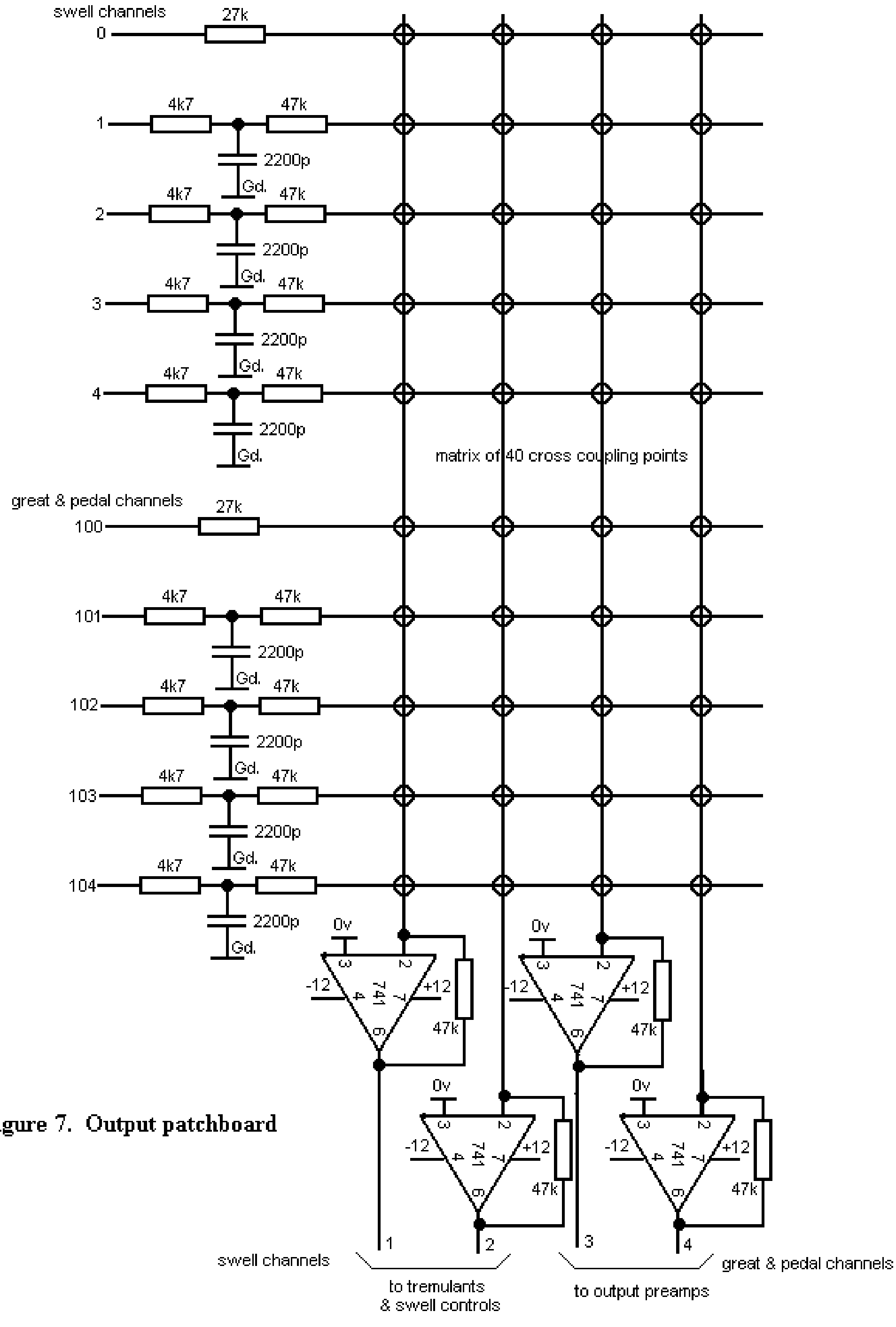


Figure 7. Output patchboard

Figure 8. Output preamplifiers

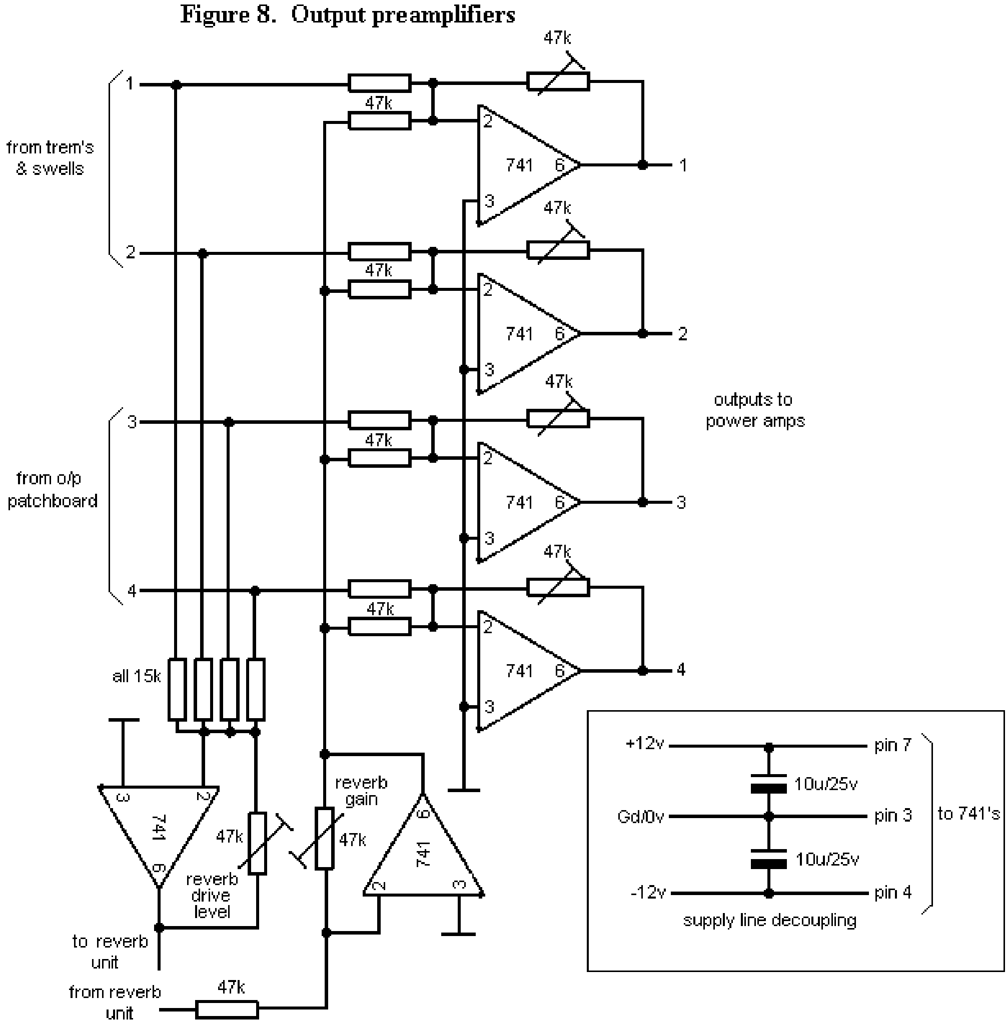
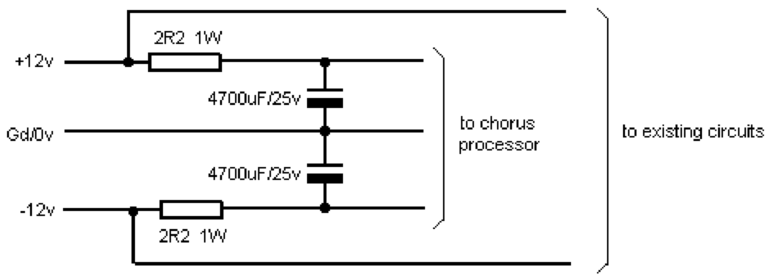


Figure 9. PSU decoupling for whole system



APPENDIX 4

A Voltage-Operated Tone-Compensated Swell Control

by C. E. Pykett

(This article first appeared in EOM 124 and 125 (February & April 1987). The MC 1495L analogue multipliers might now be difficult to obtain).

The subjective effect of a real swell box is subtle and not well matched in an electronic instrument by the usual simple potentiometer. Firstly, of course, the volume does not go to zero when the shutters are closed, but this is easily accommodated by any circuit. More important is the tone variation (change of acoustic energy distribution with frequency) which occurs as the shutters open or close. Starting from the position of maximum closure, a gradual opening of the shutters at first results in a much greater increase in the level of high frequencies than low ones. As the shutters continue to open, the volume increase becomes progressively more uniform across the spectrum. Conversely, as the shutters close the higher frequencies are attenuated more rapidly than the lower ones, with the effect becoming most pronounced near the point of maximum closure. The overall loudness variation is around 15 to 20 dB. I have not attempted to quantify the effects just described in any more detail because one organ differs so much from another that the effort is not worthwhile. It is only possible to convey the general subjective effect by a summary such as that above.

A simple circuit which provides an extremely satisfactory imitation of the sought-after effects is shown at Fig. 1. At the top end of the variable resistor, remote from the capacitor, there is no effect at all on the signal (provided it originates from a source whose impedance is low compared to R). As the slider moves down towards C there is a progressive reduction in high frequencies, which becomes more marked in terms of slider progression as it travels towards the capacitor. When the slider reaches the capacitor end of the pot the volume has not reached zero, (provided that R and C have been chosen appropriately). Subjectively, the effect is more satisfactory when a linear, rather than a logarithmic, resistor is used. We have now satisfied, in a qualitative sense, the requirements for imitating a pipe organ swell box.

The frequency response of the network, for various values of resistance in circuit, is shown in Fig. 2. The parameter k expresses the proportion of R lying between the capacitor and the position of the slider, and it varies from zero (slider at C) to 1 (slider remote from C). The corner frequency, Fo, is defined in the figure, and a value of about 160 Hz has been found to be effective, using R=l0k and C=100n. The curves show how the higher frequencies are strongly dependent on changes in k when k is small (i.e. ‘swell box closed’), but less so as k becomes closer to unity.

This simple control has precisely the subjective feel of a typical pipe-organ swell, and if this were all there was to concern us, then this article could be finished here. Some practical considerations now intrude, however.

Firstly it is not good practice to have analogue signals passing through a foot-operated potentiometer, partly because this is a component particularly prone to noise and partly because routing the signals in this way often means taking them physically away from, and then back to, other signal-conditioning electronics.

Secondly, it is really quite difficult to have a pedal, to which large forces can sometimes be applied, operate a flimsy variable resistor across the whole range of its travel; the whole range is necessary if there is to be no tone control effect at ‘swell open’ whilst enjoying the critical control over high frequencies near ‘swell closed’.

Thirdly it is often necessary to control several signal channels at once from the one pedal. Using direct control with several potentiometers in the signal lines would only multiply the other problems just outlined. Thus one is driven to consider a version of the Fig. 1 circuit which gives the same effects but is voltage-operated. Voltage control still requires a potentiometer, but the difficulties above are reduced in magnitude because the problems of noise in a worn component will not become obtrusive until a much later stage in its life. Also, it is not then necessary to operate the resistor over the full extent of its travel, and as many signal channels as necessary can be controlled from one voltage with proper design. Finally, it is not necessary to take the analogue signals on a long journey to the swell pedal and back.

An additional bonus conferred by voltage control is the ability to imitate the inertia of swell shutters. In a pipe organ having a direct mechanical linkage between pedal and shutters it is not possible to open or close the box instantaneously - the considerable mechanical inertia of the moving parts sets a limit to the speed with which the shutters can be operated. This is even more true in an instrument where the connection is through a pneumatic or electro-pneumatic system, where there is a definite time constant associated with the shutter motion no matter how fast the pedal itself is operated. In an electronic organ devoid of any attempt to simulate the inertial effect, a very rapid change in volume sounds thoroughly artificial. Voltage control can circumvent this problem if a suitable electronic time-constant is incorporated.

Readers will no doubt have their preferences for one of a number of voltage controlled analogues of the Fig 1 circuit but my chosen circuit uses analogue multipliers, since these were already in use in an earlier and less satisfactory form of swell control. The block diagram of Fig 3 illustrates the overall concept. The actual swell control operates a high quality variable resistor from which a control voltage varying from 0 to 1 volt is obtained, following normal analogue synthesiser practice. A potential of 1 volt corresponds to swell open and zero to swell closed. (Practical forms of this control will be discussed later). This passes into a control interface circuit which generates two control voltages: the ‘k’ line simply follows the input from the swell control, and the ‘1 - k’ line is the control voltage subtracted from unity. These two lines can control as many signal channels as desired, each one having the characteristics of Fig 2. The meaning of the parameter ‘k’ remains exactly as before: it is the fractional effect exerted by the swell control such that k=1 corresponds to no effect on the controlled signal, and k=0 corresponds to maximum effect. The relationship between pedal position, k and 1 - k is sketched in Fig 4.

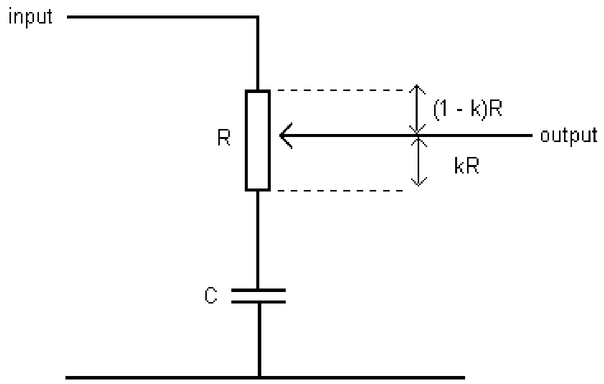
A block diagram of one signal channel appears at Fig 5. The input signal is applied to both analogue multipliers Ml and M2; in the latter case it first passes through an RC network having the same corner frequency as that considered earlier (Figs 1&2). Multiplier Ml is controlled by control line ‘k’ and M2 by ‘1 - k’. The two multiplier outputs are summed by amplifier S. Thus, when the swell control is fully open, the signal passes straight through Ml and emerges unchanged at the output; there is no contribution from M2. When the control is fully closed, Ml cannot pass any signal, and the output consists entirely of the signal passing through M2 via the RC network with its associated frequency response. At intermediate positions of the pedal, both multipliers contribute to the output. It may be shown that the transfer function of this circuit is identical to that of the simpler one in Fig 1.

The circuit diagram of a signal channel is given at Fig 6. The analogue multipliers used were low cost devices which had corresponding idiosyncrasies. The temperature variation of the offset voltages is significant, and because they operate in a non-symmetrical manner between the power rails, they are susceptible to power supply line hum and noise. The offset voltage variations can be trimmed, but in this application this was not considered worthwhile because the control voltage from the swell pedal can be adjusted such that it does not quite reach zero. This prevents multiplier Ml operating too close to its zero gain condition where offsets become most noticeable. The power line rejection problem was solved through local decoupling. Both multipliers operate with a scale factor of 0.1, the summing amplifier having a gain of 10 to restore the overall gain to unity.

All signal channels are controlled by the interface circuit shown at Fig 7 which derives the two control lines ‘k’ and ‘1 - k’ from the original control voltage. The 4k7 resistor and the 47 microfarad capacitor loading the slider of the potentiometer endow the circuit with a time constant of about 0.25 seconds, imitating the inertial characteristics of a real swell box as discussed earlier. This passive lag also acts as a filter to remove noise from the potentiometer used as the swell control.

Deriving the control voltage from the pedal is a subject worthy of comment. Fig 7 shows the variable resistor attached to the pedal being supplied with 1 volt from the 11k/1k potential divider. Obviously this can only be done where the resistor can be physically operated up to one of its limits by the movement of the pedal. Normally this is not easy to arrange with a standard rotary control because of the risk of damage from an over-enthusiastic musician. However operation between the end stops can be achieved using special purpose items marketed for the purpose. One type of control has a shaped metal cam attached to the pedal spindle which moves over and shorts a number of contact wires as the pedal is operated. The wires can be connected to a chain of equal fixed resistors hung between the 1 volt supply line and ground (but incorporating a preset at the grounded end for the reason described later). This is illustrated in Fig 8, and at least 16 wires are recommended. If a rotary control is used however, it can be mechanically connected to the pedal so that only a limited sector of the resistance track is used as in Fig 9. Preset resistors can then be used to set the voltage swing between 0 and 1 volt as in Fig 10. In any embodiment, it is advisable to include a preset at the low potential end of the potentiometer so that the minimum control voltage can be set just above zero volts. This prevents difficulties associated with the temperature dependent offsets of the analogue multipliers, as mentioned earlier.

Figure 1. Tone-compensated volume control in its simplest form



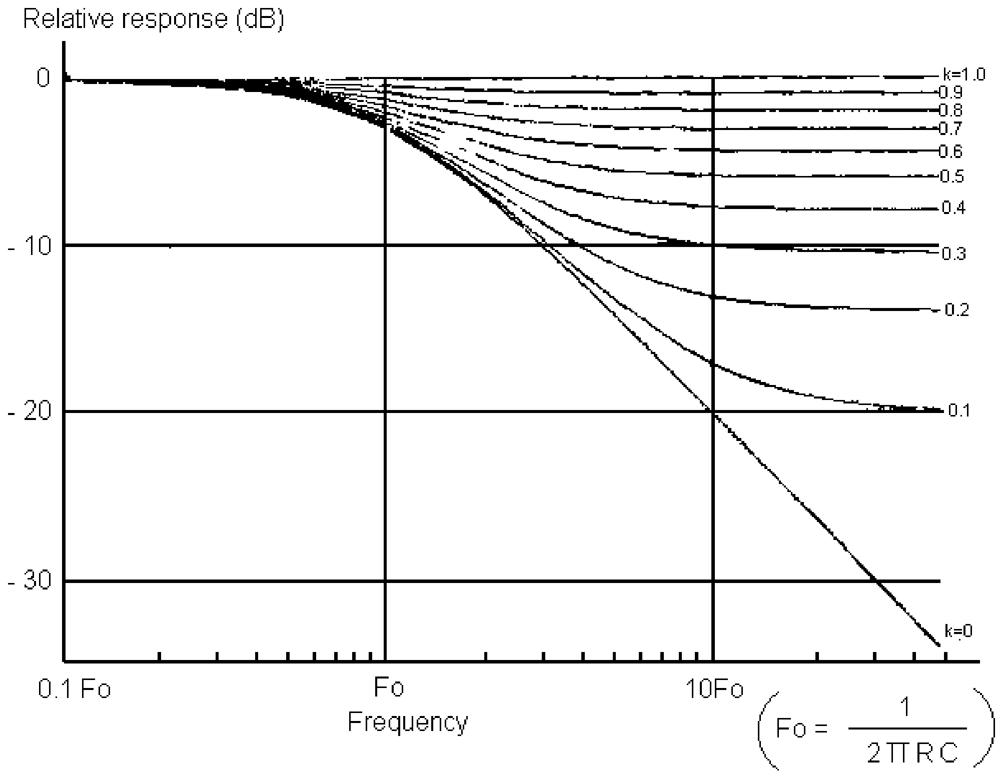
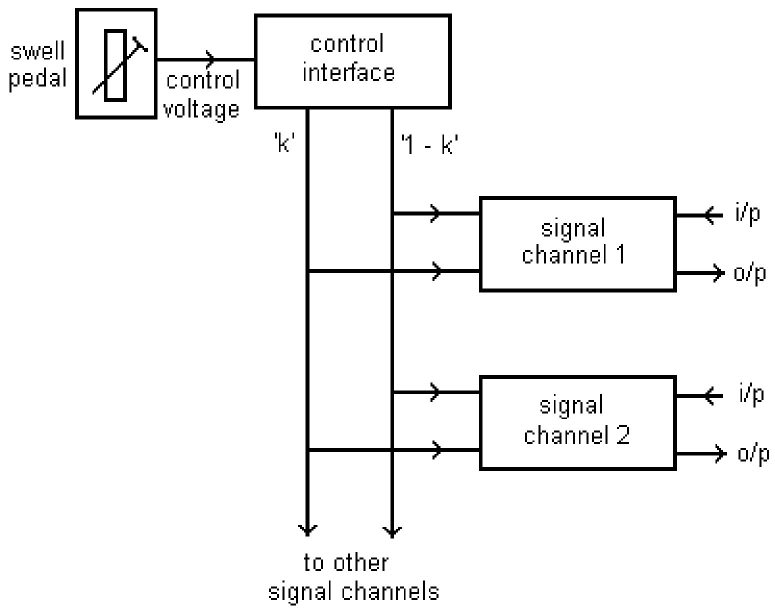


Figure 2. Frequency response of the network in Fig.1 for various values of k

Figure 3. Block diagram of total system



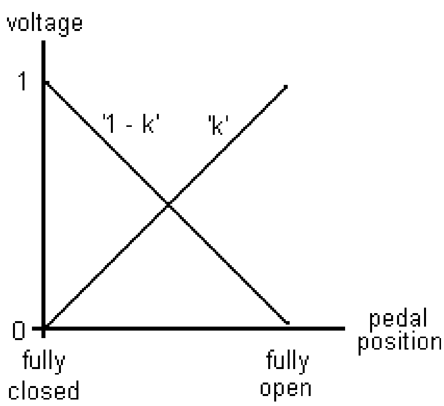
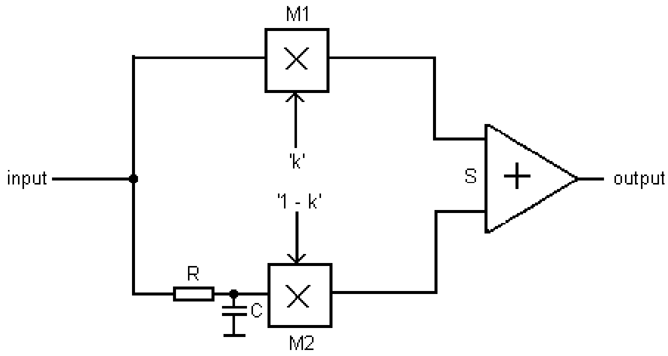


Figure 4. Relation between pedal position and control voltages

Figure 5. Block diagram of one signal channel



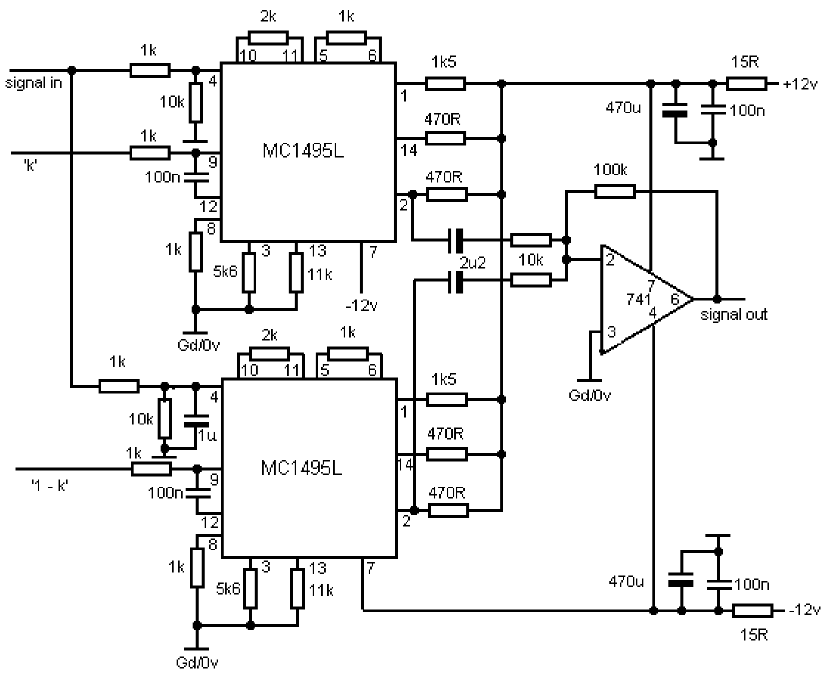
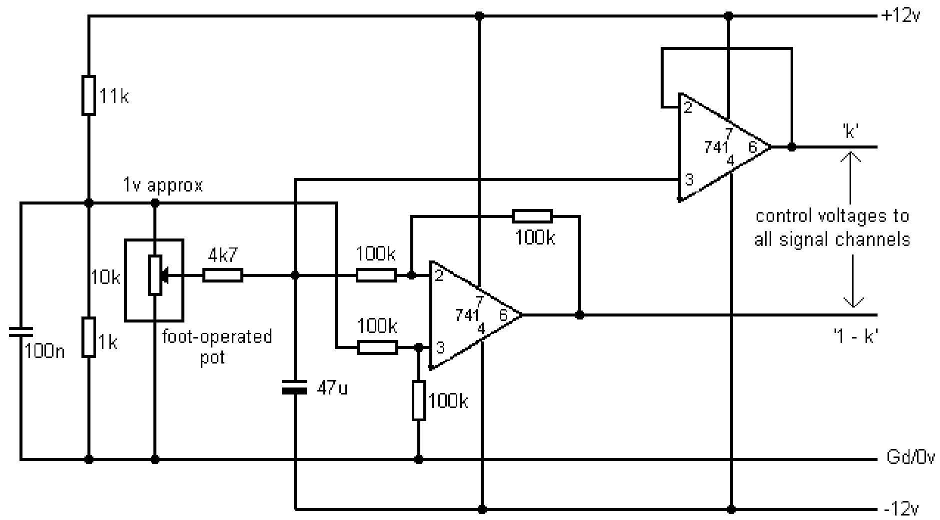


Figure 6. Circuit of one signal channel

Figure 7. Circuit of control interface



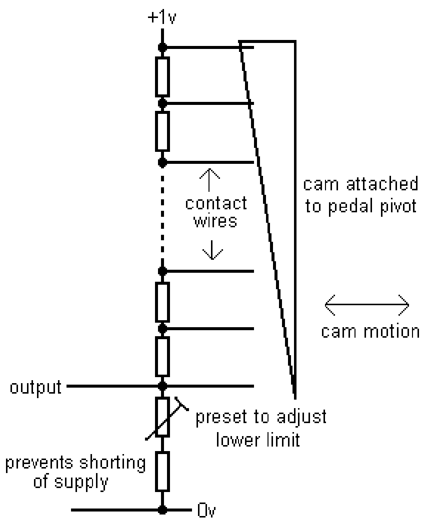
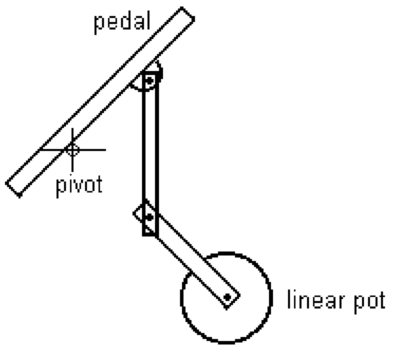


Figure 8. Illustrating use of standard organ multi-contact swell control

Figure 9. Swell pedal linkage giving an approximately linear action over a 90 degree arc (after Ryder, WW Mar 79)



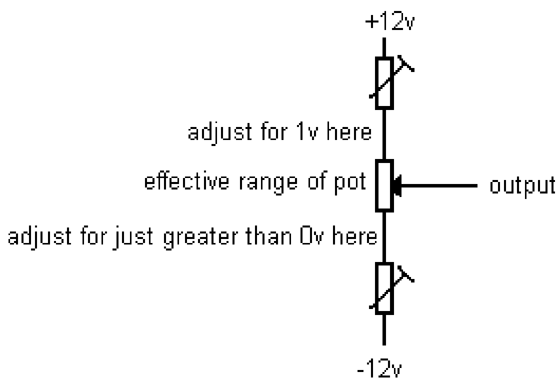


Figure 10. Method of connecting swell control in Fig 9 to obtain voltage swing from 0 to 1 volt

APPENDIX 5.

Soft Switching Of Tone Filter Outputs

by Colin Pykett

(This article first appeared in EOM 165, February 1998)

A "soft switch" is often useful in audio circuits to prevent clicks when it is operated. For analogue organs one of the most useful applications is in stop switching circuits, which generally open or close the signal path between a tone filter output and the subsequent stages leading eventually to a power amplifier.

There are many integrated circuits available which offer analogue switching capabilities. Some of the cheapest are the 4016 and 4066 quad switches. Although the 4016 is an earlier and in some ways a lower performance device than the 4066 (e.g. it has a higher ON resistance), it is attractive in that it does not seem to switch so fast, so when no audio signal is passing through it, it therefore remains quiet when operated. (The 4066 always seems to generate a slight click possibly due to breakthrough of the fast switching edge, which is speeded up inside the device). However when an audio signal is also passing through, switching transients are heard with either type because the point on the waveform at which the switch operates is seldom near to a zero crossing point. The problem is worst for large amplitude signals such as those likely to be found on loud bass stops, and in these cases it can be most unpleasant and highly intrusive when the stops are manipulated whilst playing.

Similar problems afflict even the more expensive devices. Some of these, expressly intended for switching various signal sources in hi-fi preamplifiers, have a slight inbuilt risetime delay of a few milliseconds to attenuate the clicks. However, this delay is not sufficient for low frequencies such as those found on 16 foot stops, where an even slower attack and decay characteristic is required. This is because the attack or decay needs to extend over several cycles, and when one is dealing with bottom C on the pedals at 16 feet, we are talking about a frequency of around 30 Hz. Several cycles at this frequency extend over much more than a millisecond or two.

It might be asked whether it is possible to use slowly varying switching ramps with IC's such as the 4016. One can of course try this (I did!), but the resultant switching action is still too fast owing to the internal circuitry of the IC and the characteristics of CMOS circuits in general. So it is not possible to use this approach in an attempt to slow down the action of the switch. Also it is not a good idea to use voltages which vary too slowly with CMOS as there is an appreciable current drawn from the supply at the mid-point of the ramp. If too prolonged, this could damage the device.

Any circuit for use as a stop switch has to be simple on account of the relatively large number of switches required. The OFF resistance also has to be high, otherwise breakthrough will occur. The acid test of a good stop switch is to press down all the keys in the top octave of a particular keyboard; nothing at all should be heard if all the stops are off, though in practice few organs (including commercial ones) meet this ideal. It is, however, an important test because the low level breakthrough that is often heard can be most irritating when just one or two soft stops are in use. Without going into detail, this test implies an on-to-off ratio of at least 60dB (1000 : 1), and desirably more should be aimed for.

This article describes a successful and simple stop switching circuit using a single field effect transistor plus some other essential components.

Field Effect Transistors

Field effect transistors come in two main varieties: the MOSFET and the JFET (pronounced 'jayfet'). The JFET is the earlier of the two, and although less flexible and of lower performance in some applications, it is nevertheless an extremely useful and robust device which is still an industry standard for circuit designers. In particular, it has none of the requirements for handling precautions and signal level limitations that are associated with MOSFET's, so the JFET can be handled with impunity in just the same way as a bipolar transistor. The only thing which will destroy it, again like a bipolar device, is too much current through the junctions.

The JFET (at least an n-channel JFET like the 2N3819) is very much like a thermionic triode valve in operation, and those used to designing circuits in the "old days" will find it delightfully simple to use. And with today's fad for valve amplifiers, this type of technology is not so outdated as some people might think. There are three terminals called the drain, gate and source. In triode nomenclature these correspond to the anode, grid and cathode respectively. The source is so-called because it is the source of charge carriers which carry the current between source and drain. (The charge carriers are either electrons or holes, depending on whether the source-drain channel is of p-type or n-type semiconductor. Either type can be obtained, but n-channel JFET's are the most popular). The drain is where the charge goes when it gets to the other side of the transistor, and the gate is the electrode which turns the current on or off to a greater or lesser extent. In an n-channel device the gate is always maintained at a negative (or zero) potential relative to the source; it must never go positive otherwise a gate current will flow. This is exactly like the negative grid bias which has to be established on the grid of a triode valve. There is however a major difference between the JFET and the triode in that current can flow in either direction from drain to source, whereas in a triode it can only flow from anode to cathode. This makes the JFET a very flexible device. This symmetrical bi-directional capability implies that there is no difference, electrically speaking, between the drain and the source, and to a large extent this is true. However, the substrate of the device (the block of p-type semiconductor which supports the delicate diffused regions of the transistor proper) is usually connected to the drain, which has to be borne in mind in some applications.

One of the most useful features of a JFET is that it can be used as a voltage-controlled resistor. This is shown in Figure 1. If a normal ohmmeter is connected between the drain and the source, a minimum resistance will be observed when the gate is connected to the source. That is, the gate-source voltage is zero, which is the most positive value it is allowed to have. In the popular 2N 3819 device, this minimum value is about 300 ohms. If now the gate-source voltage is made negative, the resistance increases almost linearly up to about 6 kohms, i.e. there is an approximately linear resistance variation over at least a decade of resistance. This is an extremely useful feature. Beyond this point, the resistance then increases suddenly to a very large value, and by the time the gate is 5 volts negative with respect to the source, the resistance is many megohms - the transistor is completely turned off.

Assuming the ohmeter is the simple moving coil type with an internal battery, its leads can be swapped round to reverse the polarity of the measurement. That is, the current is now made to flow the opposite way between source and drain. Similar behaviour is obtained, proving that current can indeed flow both ways through the device, but the resistance values are not the same for a given bias voltage. This is partly because the reversed current flow results in an effective gate bias different to that for the forward direction. Nevertheless, the fact that current will flow both ways means that the device can be used to control alternating currents which are symmetrical about zero volts.

By applying a relatively slow voltage ramp to the gate, the resulting resistance variation between drain and source can be made to smoothly vary the amplitude of an AC signal emerging, say, from a tone filter. By suitably selecting the time constant of the ramp, clicks in switching the signal can be entirely eliminated. Moreover, the very high OFF resistance of the device means that substantial attenuation will result when the associated stop is OFF. The chief reason for residual feed-through in these circumstances is due to drain-source capacitance, but with careful design the effects of this can be made so small that the switch is suitable for organ applications.

A number of practical considerations have to be borne in mind in an actual circuit, as shown in Figure 2. Some of these would apply regardless of the type of switch used, even a mechanical one, though this would be very "clicky". So they do not imply that features in the circuit are particularly critical in most respects.

1. The signal source has to have absolutely no DC component, otherwise clicks will be heard when the stop is operated. This is achieved by the input capacitor. The value of 1 uF may be excessive in some applications, but note that the input impedance of the switch when ON is quite low (22k) - see note 4 below. Thus the capacitor has to be chosen so as not to attenuate the lowest frequencies appreciably.

2. The 1M shunt resistor at the input is to prevent build-up of residual static charge at the drain when the switch is OFF, otherwise clicks would again result.

3. The source terminal of the transistor has to see a DC path to ground so that the gate-source bias voltage can be defined. This is achieved by feeding the source into a standard operational amplifier circuit. As indicated in the diagram, this op amp does not necessarily have to be repeated for every switch. If several stops are to share the same output channel, then their FET's (plus an associated 22k resistor) can be fed into the same virtual earth point (inverting input) of just one amplifier.

4. Notwithstanding the remarks in note 3, the impedance seen by the source terminal must not be too high otherwise feed-through via stray capacitance when the switch is OFF would become significant. Feeding the source into a moderately low impedance reduces this effect. The value of 22k chosen in the figure means that the output impedance of the tone filter must be small compared with this (i.e. the input impedance of the switch is 22k). Also the input capacitor must be chosen with this impedance in mind if it desired to use a different value from the 1 uF shown.

5. The time delay network attached to the gate produces a rising voltage ramp from - 5 volts to zero when the stop is drawn. The time constant chosen is about 50 milliseconds, which is adequate even for the lowest frequencies yet not excessive for the higher ones. When the stop is OFF note that its switch has to be CLOSED. Thus when the stop is turned off this results in a corresponding gate voltage decay to - 5 volts, resulting in the transistor going into its high resistance state.

6. Certain layout precautions should be observed to ensure minimum feed-through in the OFF state. The leads going to the source and drain terminals should not run too close to each other; this includes tracks on the PCB which should also be of minimal length. If the leads are more than a few centimetres long they should be screened, but not in the SAME multicore cable. Separate screened cables should be used for the source leads and the drain leads. Careful attention should also be paid to the grounding arrangements. The ground (zero volt) line of each switch should be run reasonably directly to a common grounding point, preferably close to the power supply zero volt connection ("star" grounding). There should not be a long and tortuous grounding route running through various other switches or signal conditioning modules. These recommendations, incidentally, are no more than the good practice which should always be followed in a large system such as an organ. They are necessary if there is to be no detectable breakthrough when stops are off, which is actually quite a tall order in view of the extreme sensitivity of the ear to such defects.

It was noted above that a JFET is bi-directional but not entirely symmetrical. The effect of this is that the waveform is turned on or off assymetrically during the time that the gate voltage is ramping up or down. Thus the positive peaks of the waveform are not attenuated at the same rate as the negative peaks, producing distortion in the signal until a steady state is reached. Generally this effect will not be audibly noticeable for two reasons: firstly the time during which the distortion occurs is only a few tens of milliseconds. Secondly, for small signals of 500 mV peak-to-peak or less, the distortion effect is small anyway. However if the switch is used to control larger signals, say of 1.5 volts peak-to-peak, a distinct though transient change in timbre might be heard as the switch turns on or off. The effect is not unpleasant, however, and some might argue that it corresponds to the starting or termination transient of a rank of pipes when their sliders are operated !

In conclusion, this switch has been used with success in analogue organs, and it has the features of simplicity as well as being effective.

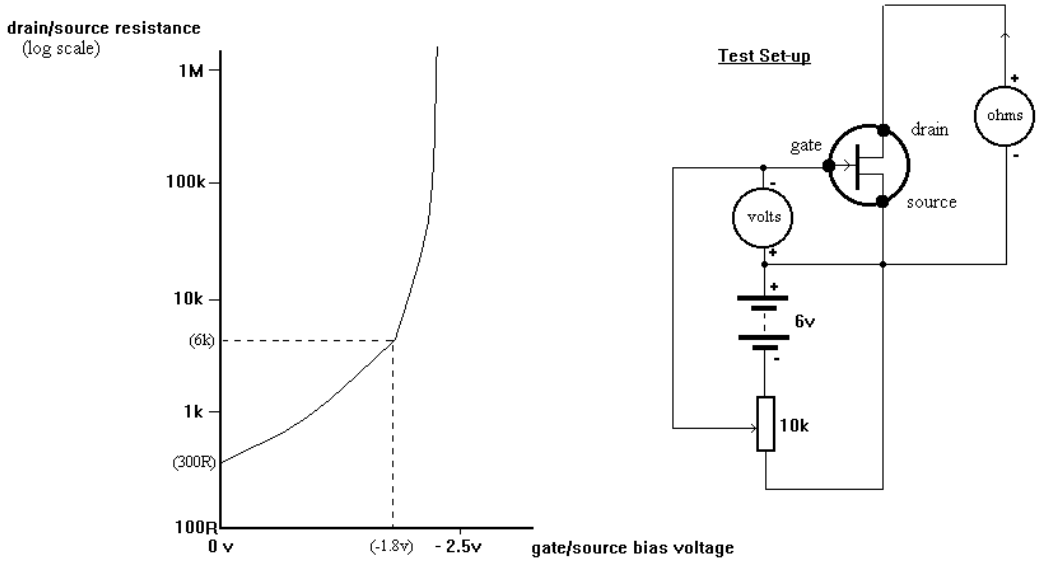


Fig 1. Resistance Characteristic of 2N3819 JFET

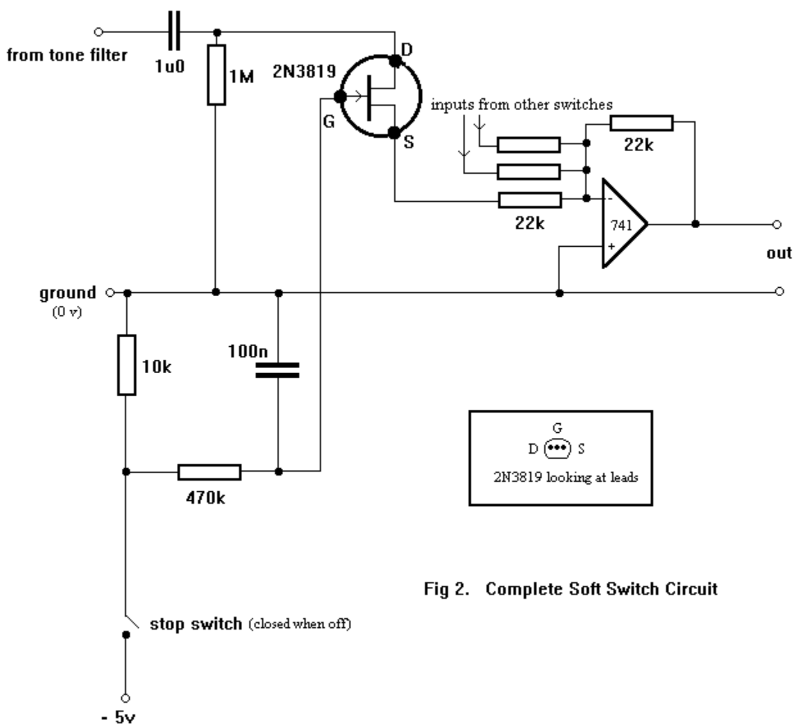


Fig 2. Complete Soft Switch Circuit