ELECTRONICS -it's easy! PART 22

More about filters

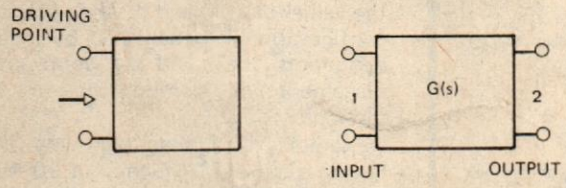

Fig. 1. Basic filter system block diagram.

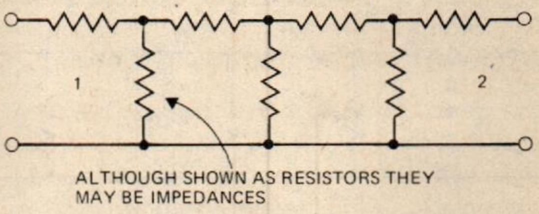

Fig.2. Ladder network.

AS WE saw in the previous section resistor-capacitor filters can only provide roll-offs of 6 dB/octave (20 dB/decade. On the other hand combinations of inductors and capacitors can provide much steeper rolls-offs and a response at the turn-over point which can be tailored to a desired shape.

The variety of LC component combinations that can be employed is great indeed and, to the uninitiated, the design of such filters can seem to be very confusing. However, circuit analysts have established design procedures which enable a filter having any practical characteristic to be designed in a logical, formalized manner. The method is based on the use of cascaded basic sections.

TWO-TERMINAL PAIR NETWORK CONCEPTS

As we have seen at various times in the course so far, filters can be circuits having just two terminals – a resonant circuit for example, or they can have two input and two output terminals – the so-called two-terminal pair networks. (The RC filter is of the two-terminal pair kind). The two different types are illustrated as system blocks in Fig. 1. Note that it is conventional to show input on the left and output on the right.

As said before many possible circuit configurations exist for filters, and the designer has to make a compromise between using a simple arrangement of many components that can be easily handled mathematically, or, a few components in a more complex network that cannot be treated by general formulae. Here we will examine the approach based on grouping numbers of simple and similar networks, to obtain the desired response, by the methods originally proposed by Zobel in 1923.

The simplest type of network is the LADDER, as illustrated in Fig. 2, the defining feature being that it has a common line. When the lower line also includes impedances (resistor elements are used to represent what are usually reactances) the network is called a LATTICE; these are much harder to design and are less commonly used. Let us examine how a ladder network is broken down into even more basic structures.

By convention the series elements of a ladder are labelled Z_1, and the shunt elements as Z_2. These elements will be either capacitors or inductors and, it is assumed that the filter is driven from, and drives into, pure resistances.

Within the ladder arrangement, shown in Fig. 2, can be seen three basic building structures – called the L section (inverted L to be absolutely correct), the T section and the π section. The three are shown separated in Fig. 3.

In Fig. 4 we see how standard T or π sections can be connected to provide the same effective ladder network. Conversely a ladder network may be subdivided into standard T or π networks by breaking the values up as shown.

The interesting and quite vital point is that the T or π stages have the same input and output impedance. That is they are *symmetrical*. The L section, however, is *unsymmetrical* in that input and output terminal pairs are not interchangeable. Two L sections in series will produce a T or a π section.

When two identical T or π sections are cascaded they are matched into the same impedance – maximum energy is transmitted and no reflections occur. Each terminal sees an image of itself, this property giving the name *image – parameter* design to this filter design method.

CONSTANT-K FILTERS

Even though a quite simple configuration has been used there can still be a wide range of combinations each with complicated mathematical solutions.

By introducing another assumption we can make some headway toward realising a wide range of characteristics with a reasonable degree of mathematical simplicity. This assumption is that $Z_1 \cdot Z_2 = R_0^2$ where R_0 is a true resistance called the characteristic resistance. (This may seem strange but the multiplication of capacitive reactance with inductive reactance yields just that). Hence Z_1 and Z_2 must be a combination of capacitor and inductor giving us

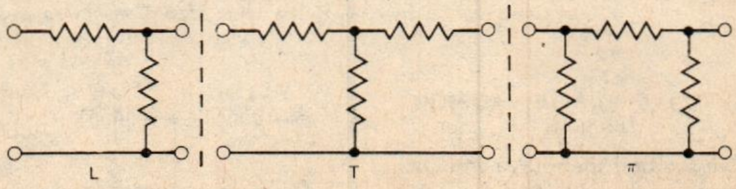

Fig. 3. The basic filter section.

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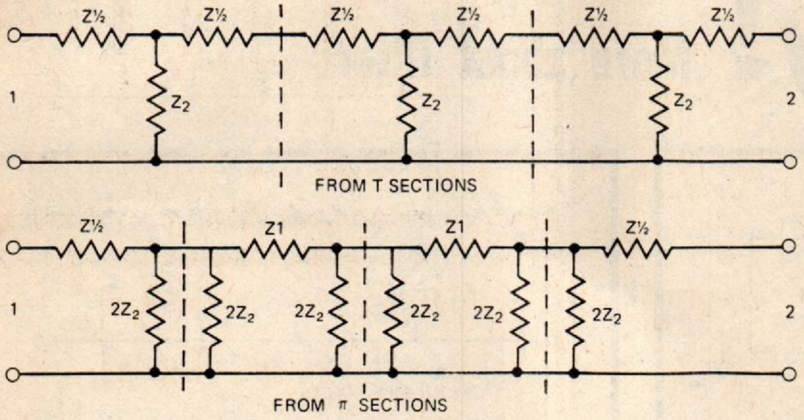

Fig.4. Building up the ladder network from basic sections.

equivalent stages with L and C proportions as shown in Fig. 5. The rule holds true for an L section provided we treat full shunt or series reactance as 2L or 2C.

The name constant-K arose from the original terminology where Zobel, in 1923, used K instead of our now accepted R_0. Filters designed to this rule are hence called *constant-K* filters.

Regardless of whether the stage is designed to be high pass or low pass — the cut off frequency will be the same, that is, at the resonance point of the LC values of the standard equivalent L section.

That is cut-off frequency $f_c = \frac{1}{2\pi\sqrt{LC}}$

For example in the T section of Fig. 6 the equivalent L section networks have L of 1 mH and a C of 0.5 microfard.

That is cut-off frequency

$$f_c = \frac{1}{2\pi\sqrt{10^{-3} \times 0.5 \times 10^{-6}}}$$

$$= 7.1 \text{ kHz}$$

Also from $Z_1 \cdot Z_2 = R_o^2$

characteristic resistance $R_o = \sqrt{Z_1 Z_2}$

However the capacitive reactance must be written as a reciprocal and in Fig. 6 this is Z_2. Hence:—

$$R_o = \sqrt{\frac{Z_1}{Z_2}}$$

$$= \sqrt{\frac{10^{-3}}{0.5 \times 10^{-6}}} = 45 \text{ ohms}$$

Thus we see that the source and load impedances used with this network must be 45 ohms, if maximum power is to be transferred, and the network is a low-pass stage having a cut-off frequency of 7.1 kHz.

If L and C were reversed the filter would have identical R_0 and f_c but it would now be a high-pass stage.

An important feature of image-parameter design is that image-matched stages can be cascaded without altering the cut-off frequencies or the characteristic resistances. Each additional stage improves the roll-off, thereby giving a powerfully reliable way to obtain the desired rapidity of attenuation without having to re-design the whole system as extra stages are added.

It can be shown that the attenuation, a, in the stop band, expressed in decibels, is a dB = 9.7 n α where n is the number of standard —T (or standard — π) sections cascaded, and α is 2 Cosh $^{-1}$ f/f_c. Cosh $^{-1}$ means the cosh function (a hyperbolic trigonometric expression) whose ratio is f/f_c. For those readers who are not familiar with this function Fig. 7 gives the relationship between values of α and frequency ratios normally encountered. Note that either f/f_c or fc/f is used depending on whichever gives a value greater than one.

The following example shows how a constant-K filter is designed to given response requirements.

The basic design formulae are:

$$L = \frac{R_o}{2\pi f_c} \quad C = \frac{1}{2\pi f_c R_o} \quad n = \frac{a \text{ dB}}{8.7 \alpha}$$

The values given at the start will be R_0, f_c, α and a dB. We need to establish, in the synthesis situation, the values of L, C and n. The necessary configuration is established by logical deduction of the appropriate placement of components in the sections.

Example: Design an high-pass filter having a cut-off frequency of 10 MHz and a signal attenuation of 100 dB at 5 MHz. The characteristic resistance is to be 50 ohms in order to match the existing system into which the filter is to be fitted.

$$L = \frac{R_o}{2\pi f_c} = \frac{50}{2\pi\, 10\, 10^6} = 0.769\ \mu\text{H}$$

$$C = \frac{1}{2\pi f_c R_o} = \frac{1}{2\pi\, 10\, 10^6 \cdot 50} = 318\text{pF}$$

To determine α

$$\frac{f_c}{f} = \frac{10.10^6}{5.10^6} = 2$$

From the chart $\alpha = 2.64$.

$$\text{Number of stages required, } n = \frac{a \text{ dB}}{8.7\alpha}$$

$$= \frac{100}{8.7 \times 2.64} = 4.35$$

We cannot however have 0.35 of a stage and therefore must use five stages to obtain at least 100 dB attenuation at 5 MHz.

The formulae for L, C are for the basic section so we have half values accordingly, giving us the circuit of Fig. 8. We could just as correctly divide the system into a π rather than a T configuration. Design of a low pass stage proceeds in just the same way.

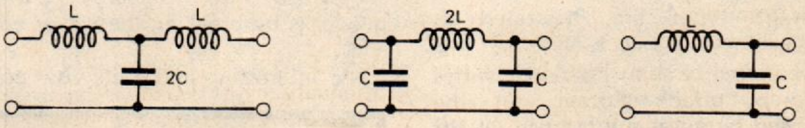

Fig.5. Convention used in this explanation of LC filter design.

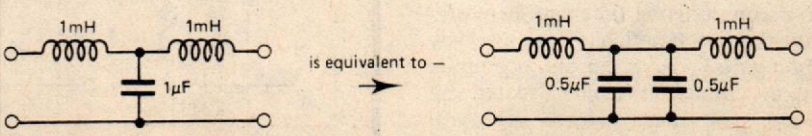

Fig. 6. A T-section filter and its equivalent L-sections.

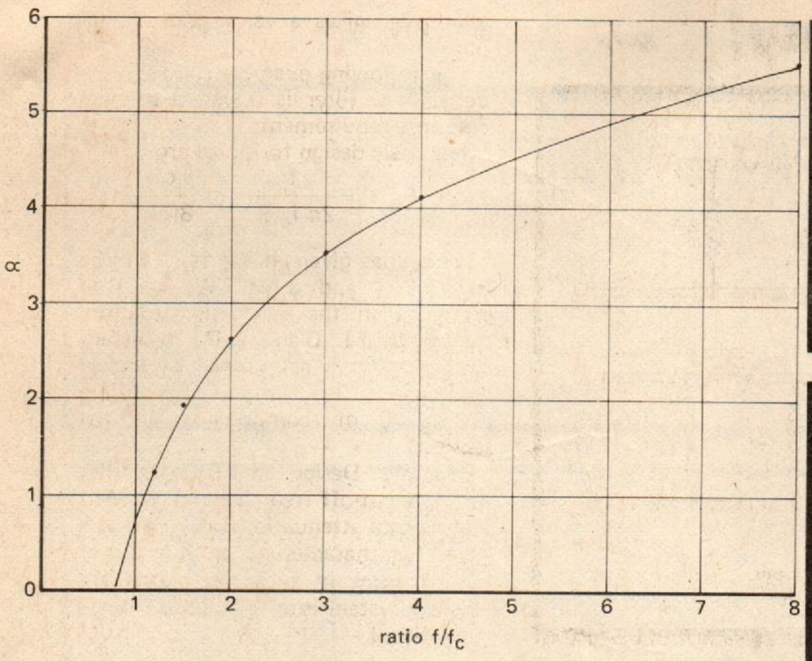

Fig.7. Chart relating value of ∝ and frequency ratio.

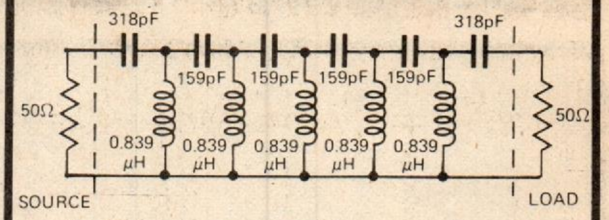

Fig.8. 5 section, constant -K, high-pass filter with cut-off of 10 MHz.

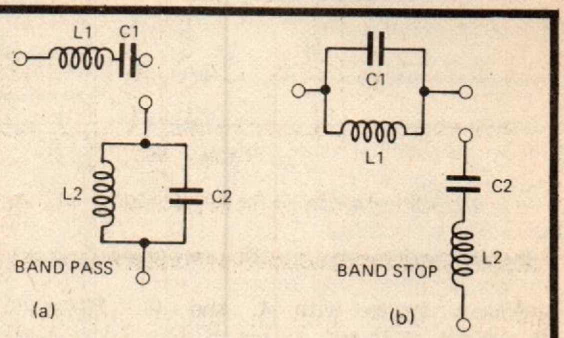

Fig.9. Basic elements of band-pass constant -K filters.

The design of band-pass and band-stop stages is more complicated going beyond the scope of this course. Suffice to say that the components in the arms now become series or parallel resonant combinations. The basic L-section for a constant-K band-pass is shown in Fig. 9a and the basic L-section for a band-stop in Fig. 9b. Readers who wish to pursue these can obtain guidance from the reading list.

M-DERIVED SECTIONS

As can be expected the simplifying assumptions made in the constant-K design, to obtain a reasonably straight forward mathematical procedure, also create practical disadvantages. The first defect is that the image impedance does not remain constant and varies in such a way that noticeable reflections occur near the cut-off points. The second defect is that the roll-off is slow just near the cut-off point: it is adequate further away from that point.

Zobel's concept to overcome this involved additional cascaded stages that, in effect, flatten out the passband response and sharpen up the cut-off point attenuation. These extra stages are called M-derived sections: one is usually added on each end of the ladder designed by the constant-K method.

We can only give an example circuit to illustrate this — Fig. 10. Although the formulae for arriving at the values are simple they must be applied with great care, the user having adequate experience in order to know the correct procedures. Again we must leave it to the reader to take this up in more specialized texts. The design of a full M-derived system requires extensive effort and training and is much more the task of a professional circuit designer than the reader for which this course is designed. The most extensive application of M-derived filters has been in communications engineering — telephones, telegraphy and multiplexed radio links. Voluminous books have been compiled that list tables giving values for chosen designs. Special computer programs have also been developed to provide automatic constant-K and M-derived section filter designs.

ACTIVE FILTERS

The basic active RC building blocks

Passive filter designs had reached their present sophistication as much as 50 years ago and in the absence of anything markedly better they continued to be the most used design until the mid 1950's. Amplification was added to make up for the attenuation that usually is experienced with passive designs.

With the introduction of reliable and less power-thirsty solid-state amplifiers in the late 1950's came the so-called active-RC filters. These combine an operational amplifier with passive RC components thereby producing filtering action more efficiently than the more obvious passive network followed by an active stage. One very valuable feature is that the effective value of, say, a capacitor can be multiplied up many times on its actual value thereby saving space and enabling designers to build circuits needing large effective values. It is also possible by active filter design to avoid the need for inductors in filter circuits. Inductors are best left out, if possible, for they are usually bulky, expensive and very lossy — they are nowhere as "ideal" as capacitors. They also are non-linear in operation and can be saturated by excessive current.

The basis of an active RC network is more often than not a reasonable quality operational amplifier set up to provide one of the following four basic circuit concepts.

1. The high gain (60 dB or more)

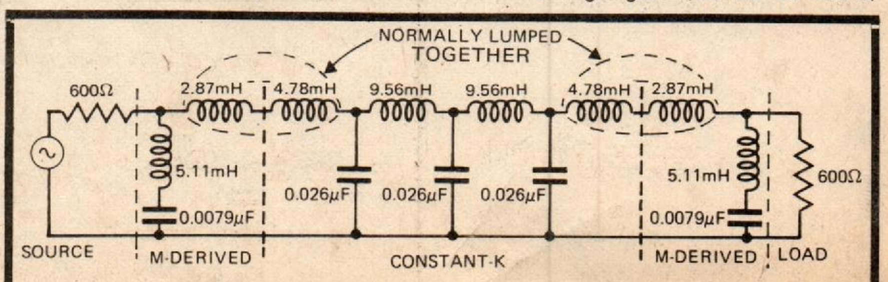

Fig.10. Example of constant -K filter with M-derived end sections.

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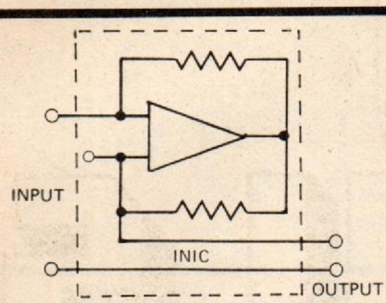


Fig. 11. Realising an INIC with an op-amp.

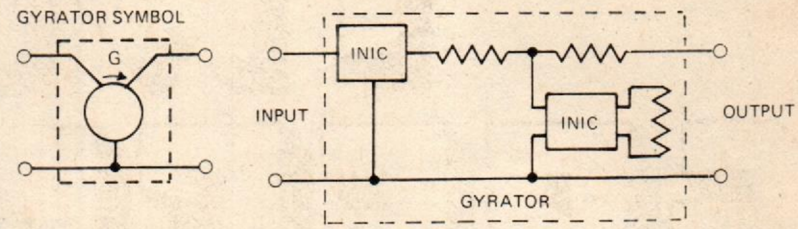


Fig. 12. The gyrator.

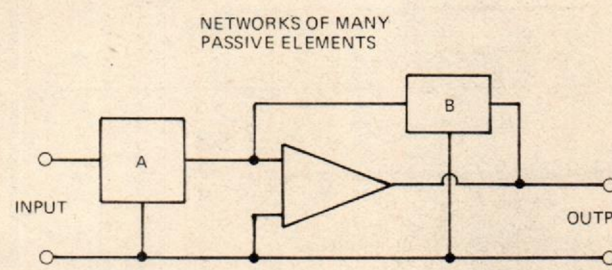


Fig. 13. Single-loop feedback active filter schematic.

voltage amplifier with close to infinite input impedance and almost zero output impedance – in short, the normal mode of an op-amp as we have discussed previously.

2. The low gain (20 dB or less) voltage amplifier, also referred to as a voltage-controlled voltage source or just VCVS.

3. The negative-emittance or negative impedance converter NIC. This is a most interesting system block for it enables positive value capacitance or resistance (that obtained with normal capacitors and resistors) connected at its input to appear as negative value capacitance or resistance at its output. It enables circuit designers to physically build circuits requiring non-physical negative capacitors and resistors. (INIC indicates an ideal current-inversion NIC, and VNIC indicates an ideal voltage – inversion NIC). A typical realisation is shown in Fig. 11.

4. The Gyrator. This is another intriguing unit for the output appears as the reciprocal of any impedance connected to its input. Thus a capacitor at its input appears as an inductor at the output. The gyrator, therefore, eliminates the need to use physical inductors and what is more, can provide more "ideal" inductors than real units. It can be realised using op-amps as shown in Fig. 12.

With these four basic possibilities available the circuit designer is rarely restricted by having synthesised a circuit needing non-physical components.

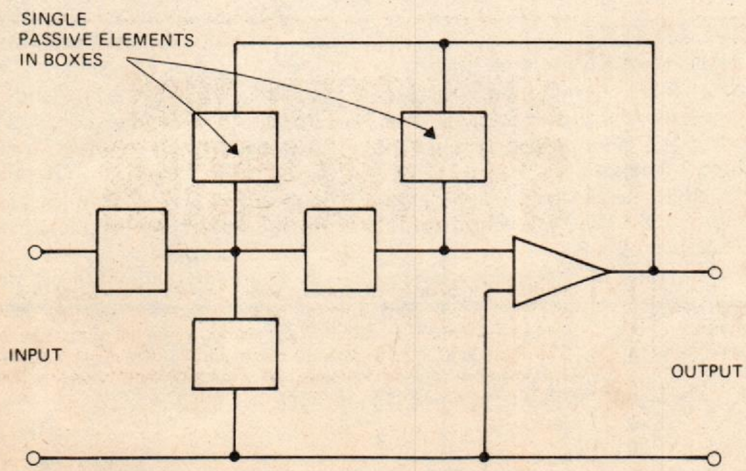


Fig. 14. The multiple-feedback-path active filter.

CHOOSING AN ACTIVE FILTER DESIGN

Given the above four system blocks it is possible to produce an incredible variety of active filters. As with advanced passive designs, few people have enough training to be expert active-filter designers. Here we can only give a guide that provides the necessary awareness of what to look for, along with words of caution as to what it is reasonable to expect from an actual active-filter design.

The voltage amplifier can be used in its simplest conceptual way with a *single-loop feed-back path* (SFP) as shown in Fig. 13 – remember how we have already seen that an op-amp integrator acts as a low-pass filter and how a notch-rejection filter, introduced into the feedback path, produces a notch-acceptance response instead.

Alternatively, we can make use of *multiple feedback paths* (MFD) as depicted in a general sense in Fig. 14, the design using minimum component count. These, somewhat surprisingly, use fewer passive elements than single-loop circuits. For this reason this form of active filter is the configuration most often used.

Our discussion about active filters will be continued next month....

ELECTRONICS -it's easy! PART 23
Last word on filters

WE CONTINUE our discussion of active filter design following our analysis of single and multiple feedback systems. The most common active filter is the multiple feedback path design shown last month. Other options open to us are to use an op-amp set up as either a controlled source with added elements — see Fig. 1, or as the negative-impedance convertor shown schematically last month. These can offer certain advantages over the voltage-amplifier design but suffer some disadvantages. NIC devices, for instance, do not give the ideal zero output impedance. Stages must be buffered to retain designed performance for example, when they are cascaded to obtain higher orders. One the good side is the small number of passive elements needed. Fig. 2 compares the four alternatives showing that no one type is exclusively the best choice.

At this stage we can only suggest that details of designs can be found in the many text books and application notes now available. Very few people would attempt (or even could) design an active filter from basic theory today. There are now available many well-prepared circuit design guides — we heartily recommend the Burr-Brown "Handbook of Operational Amplifier Active RC Networks". This contains twelve basic circuits, along with quite manageable design procedures for each, in which desired values are put in formulae to arrive at circuit values for low-pass, band-pass and high-pass requirements.

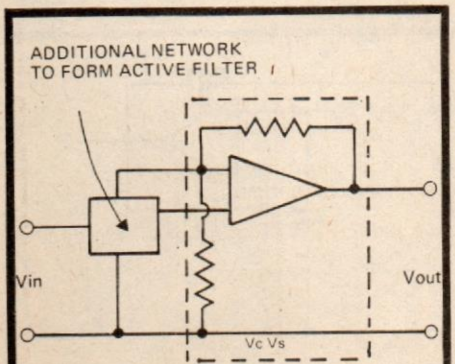


Fig. 1.

Property	Realization Technique: Infinite-Gain Single-Feedback	Infinite-Gain Multiple-Feedback	Controlled Source	Negative-impedance Converter
Minimal number of network elements	-	+	+	+
Ease of adjustment of characteristics	-	0	0	+
Stability of characteristics	+	+	-	-
Low output impedance	+	+	+	-
Presence of summing input	+	-	-	-
Relatively high gain available	+	-	+	+
Low spread of element values	+	-	+	+
High-Q realizations possible	+	-	+	+

\+ indicates the realization is superior for the indicated property

0 indicates the realization is average for the indicated property

\- indicates the realization is inferior for the indicated property

Fig. 2. Comparison table for various kinds of active filter realisations (from Burr-Brown handbook).

FILTER CHARACTERISTIC TERMINOLOGY

The ideal edge on a filter characteristic is usually a sharp "square" response with attenuation occuring instantly as the frequency passes through the corner point. It should also have a constant response level at all points in the pass-band regions. As well as the rudimentary RC filter characteristic which falls off at 20 dB/decade from a breakpoint, two other kinds of response are commonly encountered. These are Butterworth and Chebyshev responses. Both derive their names from persons who developed the mathematics involved — (Butterworth designed filters around 1930, Chebyshev developed certain mathematical theory in his study of steam-engine linkages around 1850).

The Butterworth response is said to be maximally flat (that is as flat as possible) in the pass-band region. It has the optimum constancy possible with a given number of available peaking resonances (the complex passive or active filter circuits can be regarded as a group of staggered-tuned resonating sections, each arranged to peak just aside of the others, thereby, providing a broadened response band and a reject region). Fig. 3 shows the kind of Butterworth responses obtainable. Note that each passes through the 3 dB, down half power, point. The order (a mathematical term denoting the number of resonances available) of the filter is denoted 'n' in the chart. A typical roll-off rate is 20 n dB/decade so a fourth order Butterworth response filter (which can

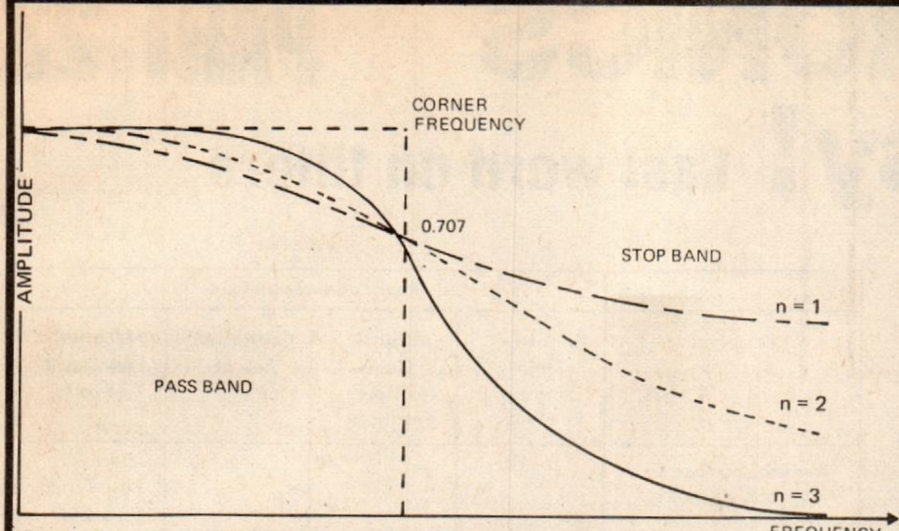


Fig. 3 Butterworth filter responses for various orders used.

be realised by either passive or active methods) will attenuate at around 80 dB/decade.

Whereas the pass-band response is reasonably constant, the rate of roll-off is not as good as can be obtained if the resonating sections are staggered differently. Other criteria of staggering the resonances can provide higher roll-off rates but only by introducing "ripples" in the pass-band response. When these individual ripples have equal amplitude across the pass-band response Chebyshev polynomials describe the shape, thus giving the name to an alternative response situation. As with Butterworth designs the higher the order the better the roll-off rate as can be seen diagrammatically in Fig. 4. The depth of ripple that can be tolerated also influences the roll-off rate — the smaller the variation that can be allowed the less the roll-off rate. (This can be readily seen by sketching in the required number of ripples of given depth at the appropriate scale).

Normally Butterworth or Chebyshev response filters will be of order 1 to 4 but higher orders are possible. These two forms are not the only sophisticated filter responses available: other mathematical criteria could be used to set up workable mathematical equations for designing other networks. These two will, however, meet most demands required and all filter design, as we have seen, is dominated by need to compromise between what is needed and what can be handled mathematically.

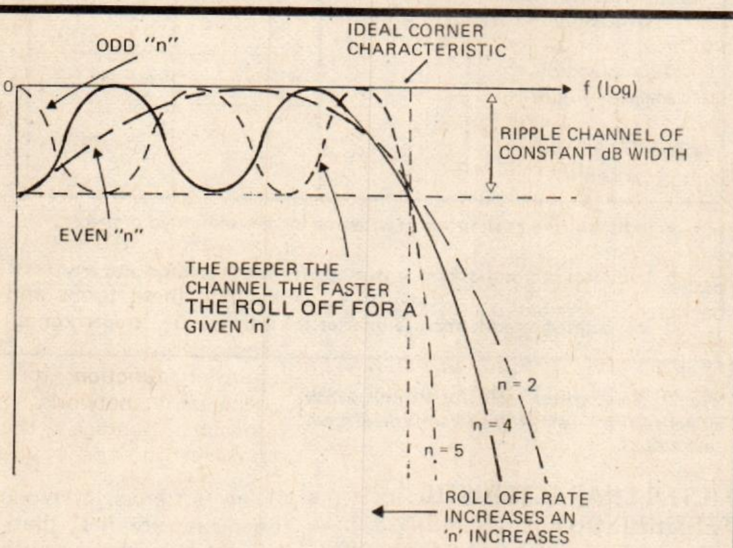


Fig. 4 Chebyshev filter responses for various orders.

PHASE SHIFT AND DELAY FILTERS

These act to provide a phase shift to a signal without selectively attenuating the frequency content. They are sometimes called all-pass filters. The amount of phase shift of practical circuits, however, usually varies with the frequency of the signal even though the amplitude response is invariant. Constant time-delay or linear-phase filters have a reasonably straight (linear) phase response as shown in Fig. 5. The so-called Bessel filter approximates this response using a workable mathematical formulation. Fig. 6 gives the general configuration of such a method realised as an active filter design.

COMPONENTS TO USE

Resistors — In non critical applications the normal 20% tolerance carbon composition resistor may be acceptable. If tighter filter characteristics are needed then one must resort to more expensive resistors such as 5% or closer tolerance carbon composition. Even better, use

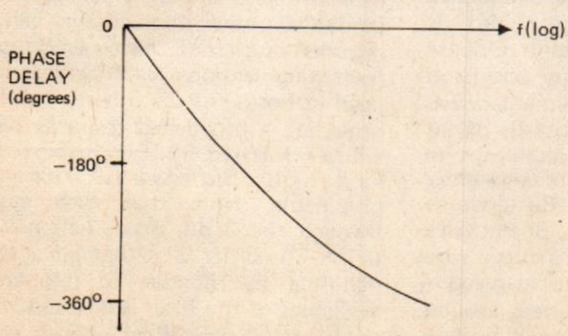


Fig. 5 Phase response of phase-linear all-pass filter.

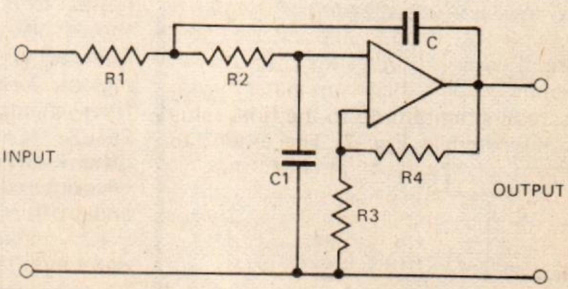


Fig. 6 General configuration of constant time-delay Bessel all-pass filter.

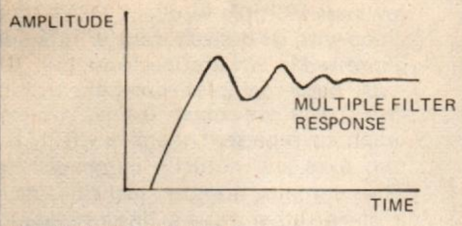


Fig. 7 Multi pole filters resonate with transient excitation.

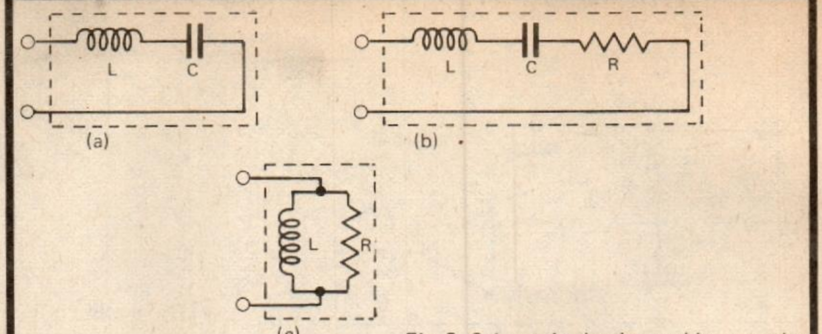


Fig. 8 Schematic circuits used in examples.

metal-film or wirewound types. It is sometimes permissible to hand choose values from wide tolerance groups in order to produce specific values, but it must not be forgotten that wide tolerance resistors often lack the same degree of time and temperature stability as the more expensive types.

Capacitors — Ceramic disc capacitors can be employed but they are best avoided. Nylon film, polystyrene and Teflon capacitors are much the better to use. When especially long-period filters are needed the capacitance value will be large. In such cases the leakage current due to losses in the dielectric is extremely critical and this rules out, in the majority of cases, using electrolytics.

Op-amp — It is easy to assume all op-amps will provide good active filters but this is not so. The main factor is a low offset current, this being especially important in long-period filters. As a general rule the more critical the need the better the op-amp should be. When op-amp filters also add gain they should have an open-loop gain at least 50 times the filter gain. Many active-filter design procedures enumerate the requirements of the op-amp.

RESPONSE TO TRANSIENTS

Filters of second order and higher invoke the characteristics of resonating circuits for their operation. In passive filters we can readily identify the inductance and capacitance; in active circuits these may not be so obvious, the mathematical expression showing that resonances do occur.

When a step change in signal is applied to a resonant circuit, the circuit 'rings', that is, the output rises rapidly but then oscillates with decreasing amplitude to the final value as indicated in Fig. 7. The extent to which a resonant circuit rings is decided by the damping provided — the higher the Q of the resonant configuration the greater the ringing effect.

It is not hard to see that higher order filters, therefore, will tend to ring more than the lower order designs when transient signals appear at their input terminals. Transients occur in practice as noise spikes, switching spikes, sudden signal appearance and departure.

THE S-PLANE, POLES AND ZEROS
(For the advanced reader)
S-Notation

The above study of filters can only act as a guide to filter selection. From there one must turn to the many articles and books available for details. To make good use of such material it is necessary to have a basic understanding of the mathematical methods used. This section is given to assist the more advanced reader. It is possible to get by without this information, provided a suitable configuration and design procedure can be located. Therefore do not be concerned if you are unable to understand this section.

Scanning through even basic, well-organised books on filter (and feedback amplifier) design the terms transfer function, s-plane, poles, zeros and root-locus will be encountered. Sadly, most books omit to provide the background explaining what this is all about. The concepts are not difficult to grasp, any confusion arising almost certainly from the number of synonomous terms used and the fact that the concepts are, perhaps, quite alien to begin with.

We have seen how reactive elements (capacitance and inductance) have apparent resistances of $2\pi fL$ for inductance and $1/2\pi Fc$ for capacitance. These terms, however, do not provide information about the phase changes produced with these reactance elements.

Electronic circuit designers use the operator symbol j (mathematicians use i) to denote a phase change of 90° hence, $j2\pi fL$ represents both the reactance value *and* the phase change. Furthermore $j = \sqrt{-1}$. For capacitive reactance the complete notation is $-j2\pi$ Fc, as the capacitor introduces a 90° phase shift of opposite sign to inductance. Resistance, having no phase shift, nor being frequency dependent is merely R. We can be a little more basic still and use ω instead of $2\pi f$. ω is the angular frequency being expressed in radians sec^{-1}. (There are 2π radians in one cycle).

When reactance and resistance are mixed we represent the value as a complex number as, for example, $R + j\omega L$. The left-hand part is known as the Real part, the other (that after j) the Imaginary part, the whole forming what is called a complex number.

Where the circuit element is only reactive the complex number representing the impedance reduces to $j\omega L$ or $-j\omega C$ for which the symbol 's' is used instead of $j\omega$. (In some books 'p' is used instead of 's'). A trap can occur here for the $-j$ of $-j\omega C$ indicates a 180° phase shift over j, not a negative quantity in the normal way. To avoid confusion we rewrite $-j\omega C$ as $1/j\omega C$ (which is valid — it comes from multiplying both numerator and denominator $-j\omega C$ by j. Hence we obtain sL and 1/sC as the shorthand way of writing inductive and capacitive reactance in which frequency dependency and phase information are both retained.

Once these terms and concepts are mastered it becomes much more straightforward to write down the transfer function for a frequency dependent network. For example, consider finding the impedance presented by a series, lossless, resonant circuit shown in Fig. 8a.

$$Z = sL + 1/sC = \frac{L\,(s^2 + 1/LC)}{s}$$

(The individual components of the expression are put on a common denominator, dividing out to get the s^2 terms with unity coefficients).

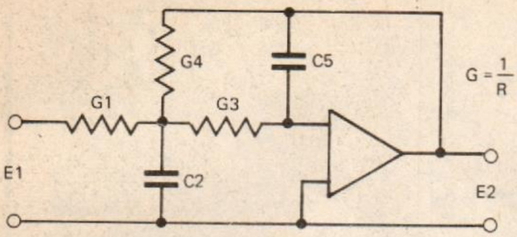
For the series resonant lossy circuit of Fig. 8b.

$$Z = sL + 1/sC + R = \frac{s^2 + R/L \cdot s + 1/LC}{s \cdot 1/L}$$

Again, for the parallel L and R circuit of Fig. 8c.

$1/Z = 1/sL + 1/R$ from which

$$Z = \frac{R \cdot s}{(s + R/L)}$$



$$\frac{E2}{I1} = \frac{-G_1 G_3}{S^2 C_2 C_5 + SC_5 (G_1 + G_3 + G_5) + G_3 G_4}$$

Fig. 9 Typical low-pass active filter with its transfer function in s-notation form.

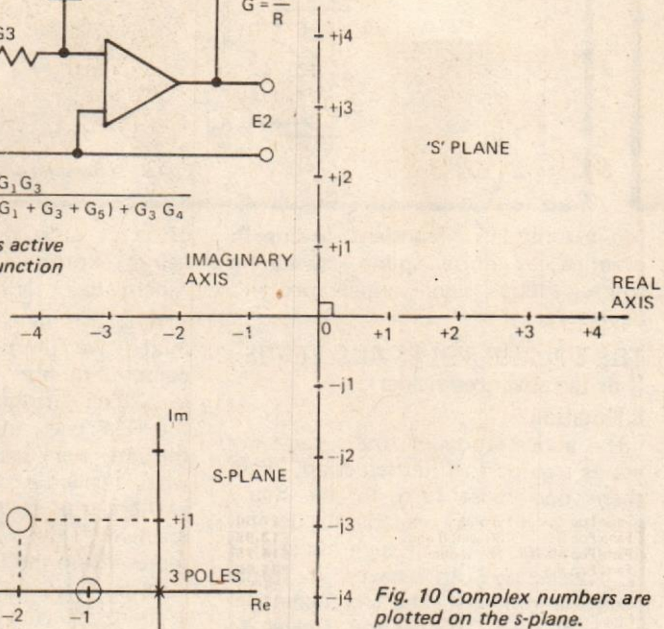


Fig. 10 Complex numbers are plotted on the s-plane.

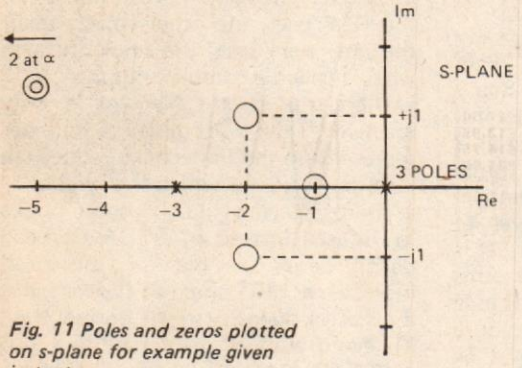


Fig. 11 Poles and zeros plotted on s-plane for example given in text.

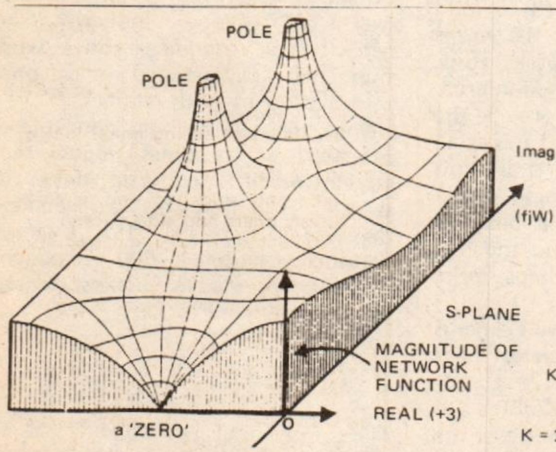


Fig. 12. *Topographical representation of poles and zeros in s-plane.*

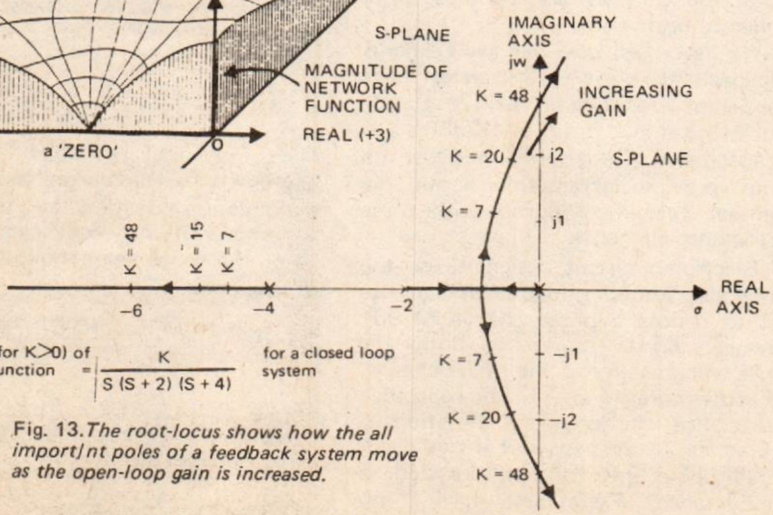


Fig. 13. *The root-locus shows how the all import/nt poles of a feedback system move as the open-loop gain is increased.*

It is these forms of expression that are quoted in circuit design books. The form of expression is not restricted to two terminal networks — it applies for all frequency dependent reactive networks. Fig. 9 gives the circuit of a low-pass multiple feedback active filter along with its derived transfer function expressed in 's' notation form.

As these complex numbers possess two parts we must use a two-axis graph to represent them in which the two axes are mutually perpendicular. Thus complex-number quantities need a plane rather than a line to depict a unique number. This plane is known as the s-plane (see Fig. 10). The two axes are usually labelled Re, R or σ for the Real axis and Im, I or $j\omega$ for the Imaginary axis, each pair being used respectively.

POLES AND ZEROS

We have seen above how a network of passive elements (active designs also apply) produces a mathematical expression in terms of s notation. As s merely represents $j\omega$ and j denotes only phase information we can, whenever s appears, substitute ω (or 2π f) to see how the expression varies in magnitude with varying frequency.

Consider the case where a function is given by the numerical example:

$$\frac{(s+1)(s+2+j1)(s+2-j1)}{s^3 (s+3)(s+5)}$$

When s = −1, −2 -j1 or −2 + j1, the numerator becomes zero for one of the bracketed terms becomes zero. Hence at each of these frequency values the expression becomes zero. We say it has 'zeros' at these points. Zeros also exist when the singular s term goes to infinity in the denominator. When s = 0 (three times, as it is from s^3 = s.s.s.), −3 or −5, we get the reverse situation for at all of these values of ω the denominator goes to zero making the function rise to infinity. These frequency points are called 'poles'.

Thus the poles and zeros express the peaks and hollows of the function. The position of these can be plotted on the s-plane diagram as shown in Fig. 11. 0 is used for zeros, a cross X for poles. In realisable networks there must be as many poles as zeros — including those at zero and infinity.

Another way to imagine the network characteristic is to draw a topographical representation giving relative height to poles and zeros on the s-plane placed horizontally as shown by the example of Fig. 12. This makes the terms poles and zeros more meaningful in a physical sense.

In the numerical example we avoided, in that case, using a quadratic or higher order term such as $s^2 + 4s + 5$. When these are encountered they must be factorized by finding the roots of the expression — giving the two terms s + (2 + j1) and s + (2 − j1) in this case. These are the individual roots, i.e., poles and zeros, of the expression. Note that quadratic elements involving an Imaginary part form mirror image pole or zero pairs — called a conjugate pair. If these are lossless (no Real part) they lie on the imaginary axis, if lossy (with Real part) they will be displaced out into the s-plane depending upon the resistive value. Positive values of resistance result in displacement into the left-hand plane, negative resistance gives poles or zeros in the right-hand plane, these halves being denoted LHP and RHP respectively.

Mathematics of complex numbers show that resonant systems with roots lying in the LHP are stable systems, their oscillations die down because to be in the LHP they must contain resistive damping. If the roots lie on the Imaginary axis itself the system is marginally stable — transients will undoubtedly create unstable situations at times even though the system is not absolutely unstable. Note that this situation only arises if the resistive component occurs as negative resistance — oscillators create this condition by the use of an active element.

ROOT LOCUS

When considering the behaviour of feedback systems, such as amplifiers, controllers and active filters, it is highly valuable to plot the changes in position on the s-plane of the closed-loop poles of the system transfer function as the open-loop gain changes. The path traced by the movement of the poles in this way is called the root-locus. These are often referred to in amplifier and other feedback-mechanism designs and it is, therefore, helpful to at least appreciate what they are. It is, however, not a simple matter to produce them from an original expression; lots of experience is vital.

By way of example the root-locus for a relatively simple transfer function is given in Fig. 13. This tells us that an open-loop gain in excess of 48 places some of its poles in the RHP establishing an unstable situation. The value of the root-locus is that we can "see" the behaviour of the system as the gain is increased and, more importantly, what we should do to the position of the poles most influencing an unstable situation. By altering the transfer function we can place the locus in more favourable situations. This is done by altering original component values where possible or by adding other networks that reduce the effect of the dominant poles — those lying close to the RHP. ●

Next month we look at digital electronics.

Digital filter design

Programming a microprocessor to act as a digital filter

by V. J. Rees, M.B.E., M.A., B.Sc., D.U.S., *Army School of Signals*

This article has three aims. The first is to answer in simple terms the question, "What is a digital filter and why use it in place of a classical filter?". The second is to show one way in which filters can be designed; the process is unfortunately rather mathematical, but even if you do not wish to follow the mathematics in detail you should grasp fairly quickly the philosophy behind it. The third aim is to show that digital filters actually do what is expected of them; a low pass filter is designed, a triangular wave of voltage applied, and the output is seen to be what one would expect from the equivalent classical filter.

Filters are a very necessary part of communications, and have been with us for a long time. They started out as LC circuits, heavy and bulky at low frequencies; some then graduated to active types, where the inductors were discarded and replaced by integrated circuits. But they still had their limitations in modern communications systems. For one thing they were not very flexible; a different set of hardware was required for each filter and if the characteristics required changing a lot of soldering had to be done. For another communications is going "all digital"; no longer is an attempt made to preserve the electrical analogue of a signal from microphone to speaker. A few samples per second of the microphone waveform are taken and we are content to send these samples, in binary form, off down the communications link.

The recipient is left with the job of reconstituting the original signal from the samples. Typical of such systems is pulse code modulation; its advantages

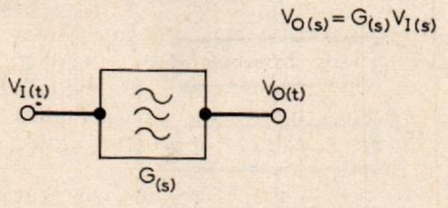


Fig. 1. Analogue filter.

are well known to communications engineers. In retrospect we are doing no more than the man who is asked to plot a graph. He will ask for a number of numerical samples from the information available so that he can plot these as points on a graph; provided he is given enough samples he can interpolate between the points and produce his own continuous function. Signals down a communications link will therefore be at only 2 levels, a 1 or a 0, representing binary numbers. This is the second reason why classical filters are not entirely happy, they are required to work in a strange world of 1s and 0s.

Two reasons have been given for considering ordinary filters unsatisfactory: their inflexibility and their dislike of a digital signal processing environment. Something which is quite happy with digits and can respond rapidly to change is the digital computer. The question now arising is whether a computer could be placed in a communications link so that it accepted samples of a signal, processed them in real time by instructions from its software, and sent them on their way "filtered." If this can be done, and it can, the process being called digital filtering, the two objections to classical filtering are overcome. Changing the frequency response of the filter is merely a matter of changing the programme; no soldering required. And of course it is happy in a digital environment, always provided it can process the sample fast enough to work in real time; the sampling theorem says we must have at least two samples per cycle of the highest frequency we transmit, say 6k samples per second for speech.

To fix our ideas on digital filters consider Fig. 1 and Fig. 2. Fig. 1 shows the classical concept of a filter; a continuously varying input signal is processed so that the voltage/time waveform of the output is different. The relation between output and input is usually expressed as a transfer function $G_{(t)}$; in Fig. 1 the Laplace transform of this function is shown. Transforms will be needed later to explain how digital filters are designed. Fig. 2 shows how a digital filter could be used to replace the classical filter. Note the requirement of some form of sampler and analogue to digital converter to produce the digital samples for the computer to work on, and the need to convert these samples back to an analogue waveform. Remember we are now working on the "points in a graph" concept.

Digital filters would have their limitations if every receiver requiring a filter had connected to it a large and expensive computer, though in some static systems the time sharing capability of the computer might make it an economic proposition if it replaced several filters. However the arrival of the microprocessor has changed all that. This is, essentially, a cheap, single chip processor programmed by an

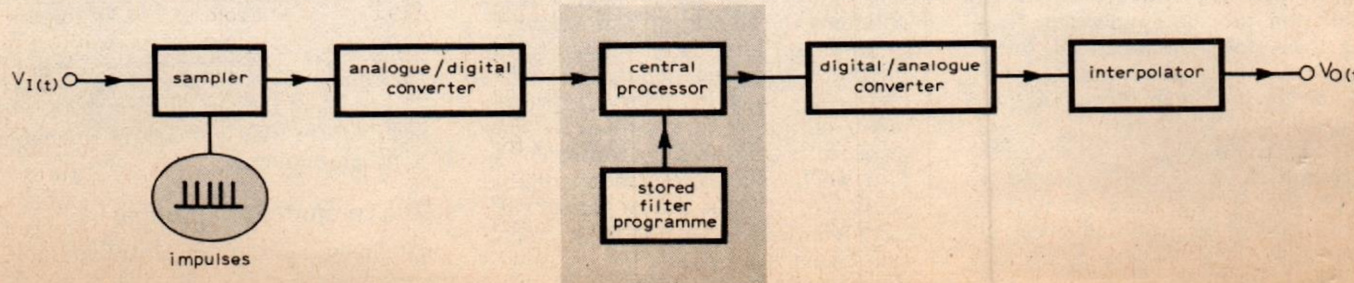


Fig. 2. Basic functions of a digital filter.

r.o.m.; it will be found within equipments fulfilling a number of functions, and used in this way the digital filter becomes an economic proposition, with of course the added advantage of changing filter characteristics simply by changing the r.o.m.

One can go a stage further, and visualise the filter becoming adaptive, that is to say looking continuously at the input signal and noise (or interference) and adjusting the filter characteristics (programme) so as to optimise the output signal to noise ratio. Perhaps the problems of electromagnetic compatibility and jamming will then assume another dimension. Certainly the future digital filter in micro-processor form looks exciting.

Digital filter design

Before proceeding to the detailed design of digital filters, a look at the problem in broader terms might help. If a computer programme is to act as a filter it must ask for a sample, process it, and send it on its way. Clearly filtering cannot be achieved without the process involving reference to previous samples; for example, a low pass filter must ensure that the rate at which sample amplitudes change is not too great to exceed the cut off frequency of the filter. Therefore the programme must hold previous samples for comparison with new samples, and we shall be looking for a programme which does this. Fig. 4 is just such a circuit and is capable of being written in computer programme form as is shown in Fig. 5. The circuit takes a sample and adds it to the sum of all the previous samples multiplied by a constant related to the filter characteristic (a) and the sampling rate ($1/\tau$); it is in fact a digital low pass filter. Fig. 4 is the heart of digital filtering. In the next section will be shown in detail how to obtain this algorithm relating input and output samples. If we accept that this can be done, filter contruction has been reduced to writing the correct algorithm to represent the filter characteristics, leaving the hardware design to the computer or better still the microprocessor designer. Of one thing you may be sure; you will be seeing a lot more of digital filters now microprocessors are with us.

And now to show that Fig. 4 is a digital filter. This is where problems arise. An elementary knowledge of Laplace transforms and impulse functions must be assumed; furthermore the modulation process involved in sampling a waveform should also be understood. You will be pleased to hear that the concept of Z transforms has been avoided.

Even if you do not follow through the mathematics, but are prepared to accept Fig. 4, the section starting "Verification of the programme" should be intelligible, and the detailed working of a low pass digital filter understood from the worked example.

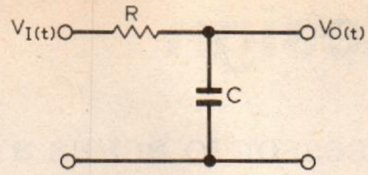


Fig. 3. Low pass filter.

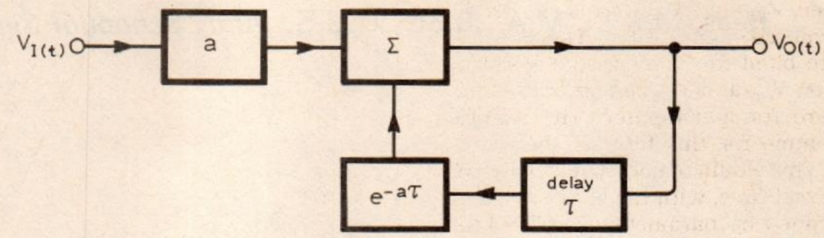


Fig. 4. Equivalent digital filter.

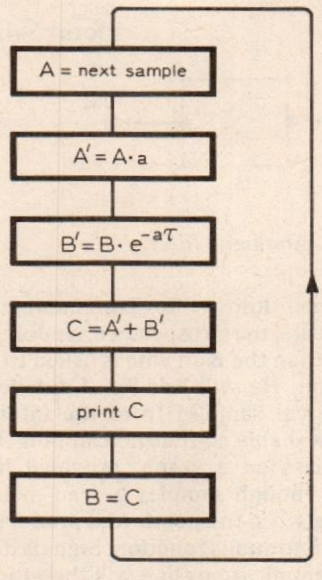


Fig. 5. Digital filter programme.

Obtaining an algorithm

Consider the transfer function of an analogue filter; in Laplace notation this is given by:

$$G_{(s)} = \frac{V_{O(s)}}{V_{I(s)}}$$

Now if $V_{I(t)}$ is an impulse function at time $t=0$, $V_{I(s)}=1$, and

$$G_{(s)} = V_{O(s)}$$

Thus the problem of determining $G_{(s)}$ can be reduced to obtaining the output of the filter when the input is an impulse. This is of course "old hat" for analogue filter designers. How does it help with the design of digital filters? The argument, which is fundamental to obtaining the algorithm, is as follows:

"In digital filtering the inputs to the filter are weighted impulse functions (actually coded in binary). Furthermore, output from the filter is only provided at the sampling times. Therefore we have constructed the digital equivalent of an analogue filter if they have identical outputs at the sampling times when an impluse function is used as an input."

An example being easier to understand than generalisations, consider the design of a digital filter to replace the simple analogue low pass filter of Fig. 3. $G_{(s)}$ can be written down by inspection, which is also $V_{0(s)}$ for an impulse input at time $t=0$; $V_{0(s)}$ can be converted to $V_{0(t)}$ by inspection. Since we are only interested in $V_{0(t)}$ at the sampling times, by obtaining this, the output of the equivalent digital filter will be obtained. Proceeding on these lines we obtain:

$$G_{(s)} = \frac{\frac{1}{sC}}{R+\frac{1}{sC}} = \frac{1}{1+sCR} = \frac{a}{a+s}$$

$$\text{where } a = \frac{1}{CR}$$

$$= 2\pi\left(\begin{array}{c}\text{upper 3dB frequency}\\ \text{of filter}\end{array}\right)$$

Solving $G_{(s)}$, by inspection, for an impulse input, we have:

$$G_{(t)} = V_{0(t)} = ae^{-at}$$

Now we are only interested in $V_{0(t)}$ at the sampling intervals 0, τ, 2τ etc. Thus $V_{0(t)}$ will be:

a volts at $t=0$

$ae^{-a\tau}$ volts at $t=\tau$

$ae^{-2a\tau}$ volts at $t=2\tau$

etc

Turning these output samples $V_{0(t)}$ into $V_{0(s)}$ will give us $G_{(s)}$ of the identical digital filter. Using the transformation

$$\mathcal{L}\left\{\delta(t-\tau)\right\} = e^{-s\tau}$$

we have:

$$G_{(s)} = a + ae^{-a\tau}e^{-s\tau} + ae^{-2a\tau}e^{-2s\tau} + \text{etc.}$$

This is a geometric progression of ratio $e^{-a\tau}e^{-s\tau}$, and the sum is given by:

$$G_{(s)} = \frac{a\left\{1-(e^{-a\tau}e^{-s\tau})^{n}\right\}}{1-e^{-a\tau}e^{-s\tau}} = \frac{a}{1-e^{-a\tau}e^{-s\tau}}$$

Thus we have obtained an algorithm, in the s plane, for our digital low pass filter. The problem remaining is to write it in such a form that it is clearly amenable to programming.

The computer programme

We have:

$$G_{(s)} = \frac{V_{O(s)}}{V_{I(s)}} = \frac{a}{1-e^{-a\tau}e^{-s\tau}}$$

i.e. $V_{0(s)}(1-e^{-a\tau}e^{-s\tau})=aV_{I(s)}$

or $V_{0(s)} = aV_{I(s)} + e^{-a\tau}(V_{0(s)}e^{-s\tau})$

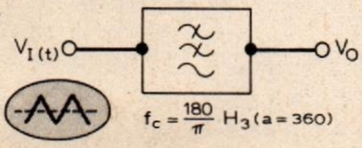
Remembering that V_I and V_0 are impulses and that

$\mathcal{L}\left\{V_{0(t)}\delta(t-\tau)\right\}=V_{0(s)}e^{-s\tau}$ it should be clear that Fig. 4 has the above relation between $V_{0(s)}$ and $V_{I(s)}$ and is therefore the required digital filter. There is no need to build this filter as the relation between $V_{0(t)}$ and $V_{I(t)}$ can be written as software for a computer. The simple programme for this filter is shown in Fig. 5. This would of course normally be run in real time, with the sampling rate determined by parameters outside the control of the filter. Intuitively it can be seen that filtering is achieved by comparing one sample with a modified version of the previous output, a process similar to delta modulation; it is ironing out the rapid changes which would be outside the filter bandwidth.

Verification of the programme

Confidence in the ability of digital filters to fulfil their purpose can best be obtained by specifying a function $V_{I(t)}$ and obtaining the corresponding $V_{0(t)}$. The programme is so simple that this can be carried out (not in real time!) by the following process:

Time (ms)	V_{in} (volts)	$a\tau V_{in}$ (volts)	$e^{-a\tau}a\tau V'_{out}$ (volts)	V_{out} (volts)
0	0	0	0	0
1	2	0.72	0	0.72
2	4	1.44	0.500	1.94
3	6	2.16	1.355	3.52
4	8	2.88	2.453	5.33
5	10	3.60	3.721	7.32
6	8	2.88	5.108	7.99
7	6	2.16	5.573	7.73
8	4	1.44	5.395	6.84
9	2	0.72	4.769	5.49
10	0	0	3.830	3.83
11	−2	−0.72	2.672	1.95
12	−4	−1.44	1.362	−0.08
13	−6	−2.16	−0.056	−2.22
14	−8	−2.88	−1.546	−4.43
15	−10	−3.6	−3.088	−6.69
16	−8	−2.88	−4.666	−7.55
17	−6	−2.16	−5.265	−7.42
18	−4	−1.44	−5.186	−6.62
19	−2	−0.72	−4.619	−5.34
20	0	0	−3.725	−3.73
21	2	0.72	−2.599	−1.88
22	4	1.44	−1.311	0.13
23	6	2.16	0.090	2.25
24	8	2.88	1.570	4.45
25	10	3.60	3.105	6.70
26	8	2.88	4.678	7.56
27	6	2.16	5.273	7.43



$V_{I(t)}$ symmetrical triangular voltage fundamental frequency 50Hz amplitude ±10V

τ 20 samples per cycle giving $\tau = 10^{-3}$ sec.

$a\tau = 0{\cdot}36$ $e^{-a\tau} = 0{\cdot}6977$

Fig. 6. Filter and input parameters.

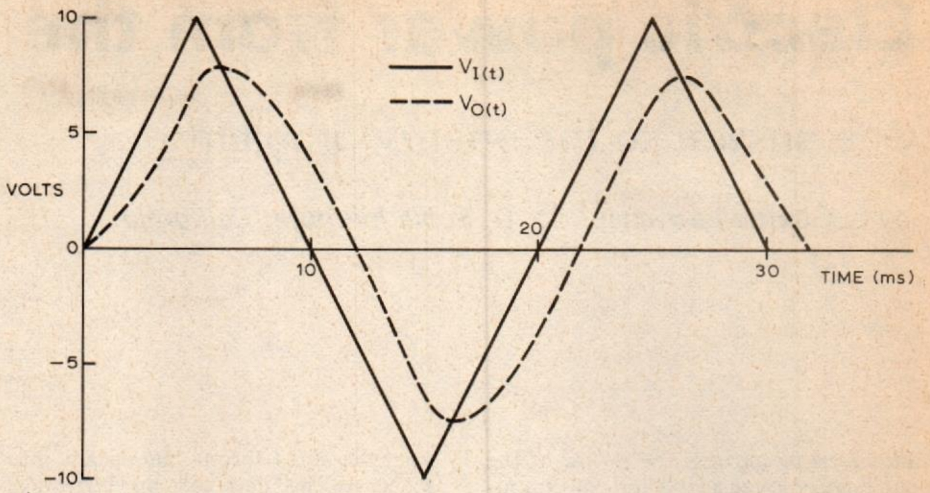


Fig. 7. Input/output signals for low pass digital filter.

(a) Draw a graph of the analogue waveform $V_{I(t)}$ which you wish to filter

(b) Decide on a sampling rate, i.e. determine τ, and obtain from your graph $V_{I(t)}$ at the sampling times 0, τ, 2τ etc.

(c) Decide on the filter 3dB frequency; this determines the product CR and a. If you know nothing about the spectrum of a sampled waveform you will need to do some further reading to understand the limitations set on the relative values of filter bandwidth, sampling rate and input signal bandwidth.

(d) Put your weighted impulse functions (samples) through the programme manually to obtain $V_{0(t)}$ at the sampling times.

(e) Multiply the output samples by τ, plot them on the same axes as the input and interpolate between them by hand to obtain the analogue output signal. In practice, interpolation would probably be achieved with a sample and hold circuit. The requirement to multply by τ needs justifying. Without doing this the shape of the output version would be correct, but the amplitude too large. It is necessary to multiply by τ because in sampling the input signal, much of the sampled spectrum lies outside the filter bandwidth, and an adjustment must be made for this, which can be shown to amount to multiplying by τ. However, in practice absolute values of output are not important; gain can be built into the system.

(f) Ask yourself whether the output you have obtained is what you would expect from the filter. This check can be carried out either by considering the Fourier series for the input, or in the case of pulse inputs by considering the application of "step functions" to RC circuits.

The above process has been tried by the author for different forms of input signal. A worked example, the results of applying the triangular wave of Fig. 6 is shown in the table and Fig. 7. The 3dB frequency of the filter and the repetition frequency of the triangular wave have been chosen so that one would expect the output to be almost entirely the fundamental compoment of the Fourier series, and this is seen to be so. It is suggested that serious readers of this paper try a square wave input (check the output by "step function" approach), and also produce the algorithm and programme for a simple CR high pass filter.

Literature Received

Microprocessor Series 8000 users' manual is now available. The manual gives a general description of the 8000 family, details of the hardware available, and assistance with interfacing and programming. The users' manual costs £5 from General Instrument Microelectronics Ltd, 57-61 Mortimer St, London W1N 7TD.

Procedures under the Health & Safety at Work Act are outlined in a short leaflet, "Regulations, Approved Codes of Practice & Guidance Literature," available free from local offices of HM Factory Inspectorate, HM Inspectorate of Mines & Quarries and HM Alkali Inspectorate.

Wire-wrap boards, designed for microprocessor chip sets, are described in a brochure from Nimrod Electronics Ltd, 85 High Street, Billingshurst, West Sussex WW401

Harris operational amplifiers, a-to-d converters and associated devices, memories, and digital i.cs are briefly described in a short catalogue, distributed by Memec Ltd, The Firs, Whitchurch, Aylesbury, Bucks WW402

Intersil tell us that they have published a guide to their range of discrete semiconductors, which contains application guidance and a cross-indexed list of devices. Intersil Inc., 8 Tessa Road, Richfield Trading Estate, Reading, Berks. WW403

A book entitled "Thick-film Conductor Survey" is now obtainable from the Electrical Research Association. The book seeks to rationalize the huge amount of commercial data and presents information on formulation, characteristics, product lists, detailed data in various categories of characteristics, prices and souces. The material presented is international and is claimed to be independent. The 200-page book costs £45 (£41 to Association members) and is available from ERA at Cleeve Road, Leatherhead, Surrey KT22 7SA.

Static smasher

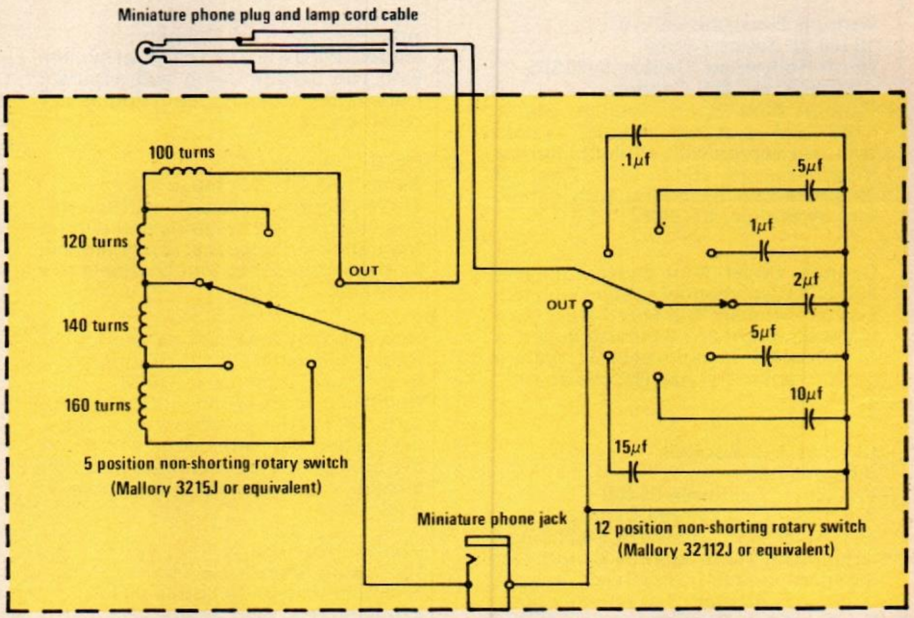
Snap, crackle and pop may be okay in your cereal bowl, but it sure is a pain in your radio. Here's an easy-to-build static filter that can cut the noise level down to size.

Electrical noise—it's the scourge of am radio. Whether it's from an automobile ignition system or a vacuum cleaner, or even mother nature, this background noise can make reception of weak signals a nightmare. That's why most CB radios have automatic noise limiters built-in. But even with a noise limiter, a weak signal can still get buried in the noise.

Generally, the noise you're trying to eliminate is at relatively high audio frequencies—above those of the signal you're trying to copy. Base station operators are sometimes bothered by 60 or 120 Hz hum as well. What's needed is a filter that is smart enough to pass through the signal you're trying to copy while at the same time eliminating both lower and higher frequency signals. And that's just what the Static Smasher does.

The Static Smasher consists of a capacitor and an inductor connected in parallel. The resulting circuit is called a bandpass filter because it passes only a relatively narrow band of frequencies. Unlike many other bandpass filter circuits you may have seen, the Static Smasher gives you a total of 28 different passbands to choose from.

Reducing or eliminating electrical noise is not without its cost, however. By restricting the bandwidth through which the signal passes, you'll reduce the fidelity of the signal you hear. But that's a small price to pay for the ability to copy signals that would otherwise be buried under the noise.



Just plug it in

The Static Smasher filters the signal going to your speaker or headphones, so you won't have dig into your radio. If you plan to use the filter with more than one radio, you can build it so that it plugs into the radio, and the speaker plugs into it.

The circuit is really nothing more than a pair of rotary switches with a tapped coil and a handful of capacitors, all of which can be mounted directly on the switches. So, you can build the Static Smasher in just about any enclosure you have handy.

If you plan to use the filter only with your base station, and have no enclosure on hand, the Radio Shack 270-260 wood-grained project cabinet should be ideal. For strictly mobile use, any aluminum box will do.

The Static Smasher is a snap to build and shouldn't take more than a few hours to complete. The most difficult part of the project is winding the coil. It's made by wrapping number 18 enameled wire, sometimes called 18 guage magnet wire, around a ¾-inch wooden dowel.

As you wrap the wire around the dowel, you'll have to keep track of the times you've gone around it. Each time the wire completes one trip around the dowel, you've added one more *turn* to the coil.

After you've put 100 turns of wire on the dowel, bring the wire away from the dowel to a distance of about four inches. Then, bending the wire back on itself, resume wrapping the wire around the dowel. After you've added another 20 turns, bring the wire away in the same manner. Repeat the procedure again after another 20 turns have been added, then complete the winding by adding a final 20 turns. The result will be a 160-turn coil with taps at 100, 120 and 140 turns.

As you wind your coil, keep the wire confined to a 1-½ inch section of the dowel. You can slip two pieces of cardboard onto the dowel and space them an inch and a half apart to help you hold the dimension. Or you can cut the dowel to exactly 1-½ inches and glue cardboard to each end.

Easy to use

The Static Smasher requires no external power. Just plug it into the speaker line and you're in business. The front panel switch controls let you choose between seven different capacitors and four different inductances. You'll find that the best combination of capacitor and inductance will vary depending on the noise and the signal you're copying.

The Static Smasher is an effective filter for reducing electrical noise interference in am radios. It is not the complete answer to the problem. You can further improve your reception by making every effort to reduce the cause and the amount of noise being generated in your car and electrical appliances. ■

If interference appears, complain to the FCC, documenting it as well as possible. They'll politely tell you that there's nothing that they can do, but if enough modelers complain in a given area and time, they might feel or become inspired to find a way. One thing is dead certain: They won't do a thing for people who DON'T complain!

Two and Three Channel Radio Control Equipment

Listed in alphabetical order

ACE R/C Inc.
Box 511H
Higginsville, Missouri 64037
Attention: Mr. Paul Runge
Digital Commander R/C kit—capability of three channels. Two servos included. Transmitter has a dry 9 volt battery. NiCad receiver battery pack with charger. Kit price $119.95 with standard size (ACE/Bantam) servos. $124.95 with Dunham D-5 servos.

Cannon Electronics Corp.
13400-26 Saticoy Street
North Hollywood, California 91605
Attention: Mr. Bill Cannon
Cannon Mini Sport System (Model 810B-22A) is a two channel system with two servos with all dry batteries. NiCads can be added later. Five full channels can be added with simple factory update. $119.95.

Cannon Super Mini System (Model 820-22A) two channel system with two servos (Dunham D-5 Micro size). Two channel weight of 3.2 ounces includes a 100 MAH NiCad receiver battery pack. Transmitter has NiCads as well. $199.95.

Charlies R/C Goodies
P.O. Box 192
Van Nuys, California 91408
Attention: Mrs. Charlie Cannon
Essentially a kit version of the Cannon equipment. Three channel kit with two standard servos (one already assembled). Transmitter uses nine dry alkaline batteries. Receiver pack is NiCad with a charger. $109.95 with standard servos or $124.95 with Dunham D-5 micro servos.

Cox Hobbies Inc.
1505 East Warner Ave.
Santa Ana, California 92702
Attention: Mr. Lee Renaud
All systems mentioned here have two servos and use eight dry alkaline batteries in the transmitter and four more alkaline batteries in the receiver pack. NiCad packs can be purchased at a later time.
Model 8021 has a wheel control and throttle lever expressly set up for R/C cars and boats. (27 MHz frequencies only)
Model 8020 has two stick (separate) transmitter. For the Mode I flyers.
Model 8022 has single stick/two axis transmitter control. Best suited for model airplane control.
Model 8031 has single stick/two axis transmitter with a third channel available for throttle control.
Prices range from $99.95 to $109.95.

EK Products Inc.
3322 Stovall Street
Irving, Texas 75061
Attention: Mr. Bill Haga
Nimbus Sport Two is a two channel system with two servos and a single stick/two axis control using all dry batteries. Servos are the miniature EK-SM model. $129.95.

Futaba Industries U.S.A.
630 Carob Street
Compton, California 90220
Attention: Mr. York Daimon
Model FP-2GA is a two channel system with two servos using two separate transmitter sticks (Mode I) and all dry batteries. $99.95.

Heath Company
Benton Harbor, Michigan 49022
Three-channel system kit with two servos using NiCad batteries and plug-in frequency modules that is expandable by the kit builder to four channels at a later time. $179.95

Hobby Lobby International Inc.
Route 3 Franklin Pike Circle
Brentwood, Tennessee 37027
Radio is manufactured for Hobby Lobby by EK Products and is a three channel system with two servos with 9 volt dry battery in the transmitter and four alkaline dry batteries in the receiver pack. $120.

Hobby Shack Inc.
18480 Bandilier Circle
Fountain Valley, California 92708
Attention: Mr. Paul Bender
Aero Sport Two system (actually manufactured by Futaba Industries) is a two servo system with all dry batteries. NiCad batteries can be added to receiver at a later time. A great buy at $75.

Kraft Systems Inc.
450 W. California Ave.
Vista, California 92083
Attention: Mr. Marty Barry
Model KP-2A Sport Series is a single stick/two axis control system using all dry batteries. Extremely reputable service is available throughout the country. $130.

Model Rectifier Corp.
2500 Woodbridge Ave.
Edison, New Jersey 08817
Attention: Mr. Frank Ritota
Model 772 is a two channel system with a two control stick (Mode I) Transmitter that comes with dry batteries. NiCads can be easily added later. Receiver is also dry powered but NiCads can be purchased as an option later on. $110.

Millcott Corp.
1420 Village Way, Unit E
Santa Ana, California 92705
Attention: Mr. Hugh Milligan
Single Stick Specialist Three is a three channel system using three servos and complete with NiCad batteries and charger. Has a special built-in mixer control for use with "V" tail aircraft. Highly specialized! $275.

Pro Line Electronics Inc.
10632 N. 21st Ave.
Suite 11
Phoenix, Arizona 85029
Attention: Mr. Jerry Bonzo
Three channel Competition Series Model PLN-3-0 has three servos plus full NiCads. $300.

Royal Electronics Inc.
3535 S. Irving
Englewood, Colorado 80110
Attention: Mr. Sid Gates
Kit system components which must be purchased individually. A two channel transmitter kit with NiCads is $75. A two channel receiver kit (less connectors) is $22. Servos and battery packs available to suit your needs.

RS Systems Inc.
5301 Holland Drive
Beltsville, Maryland 20705
Attention: Mr. Frank Goodwin
Model RS-3-S0 is a three channel system with two servos and full NiCads. $235.

Tower Hobbies Inc.
P.O. Box 778
Champaign, Illinois 61820
Three channel system with two servos actually manufactured by Kraft Systems. Transmitter uses a 9 volt dry battery but can be converted to NiCads later on. Receiver comes with NiCad pack and charger. $120.

World Engines Inc.
8960 Rossash Ave.
Cincinnati, Ohio 45236
Attention: Mr. Dave Brown
Expert Series two channel system has single stick control and comes complete with dry battery system for $135 or with full NiCad system for $180.

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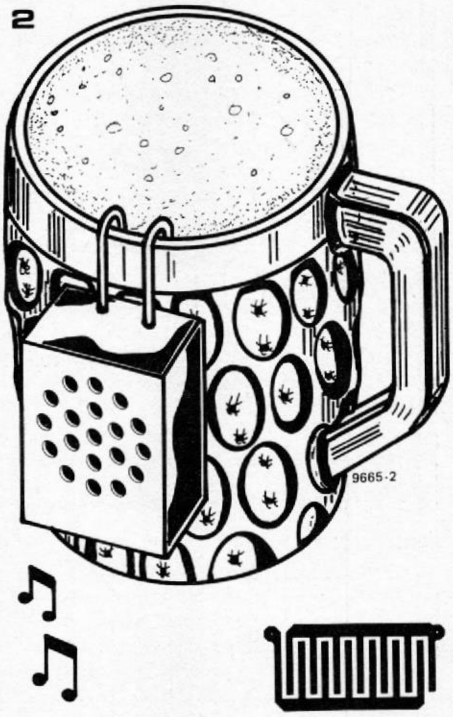
24 liquid level indicator
25 aerial booster
26 variable slope filter

24 This circuit was originally intended as a water level indicator for use by blind persons, to give an audible indication when a cup, bowl or other container was full. It will function with any liquid that will conduct electricity, such as beer, tap water, tea, milk. It will, of course not function with distilled or de-ionised water. The circuit has other applications such as a rain sensor (when used with a suitable probe).

The circuit is extremely simple. The input of N1 is normally held low by a 1M resistor. When the probes are immersed in a conducting liquid the input of N1 goes high, so the output goes low and the output of N2 goes high, enabling the astable multivibrator N3/N4, which switches T1 and T2 on and off to produce a tone from the speaker. An open collector transistor output is also provided to drive a relay or other circuit. Probe construction for level sensing and for rain sensing are shown in figure 2. The level sensor probes should preferably be made of stainless steel wire for ease of cleaning, and the circuit housing should be watertight in case of accidents.

2



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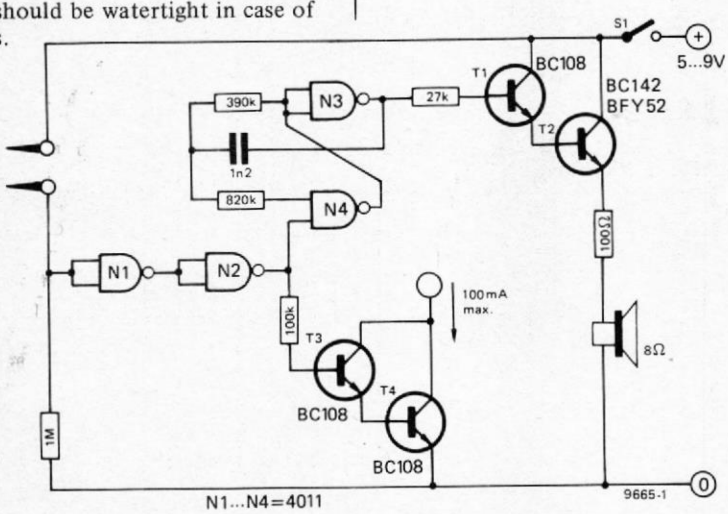
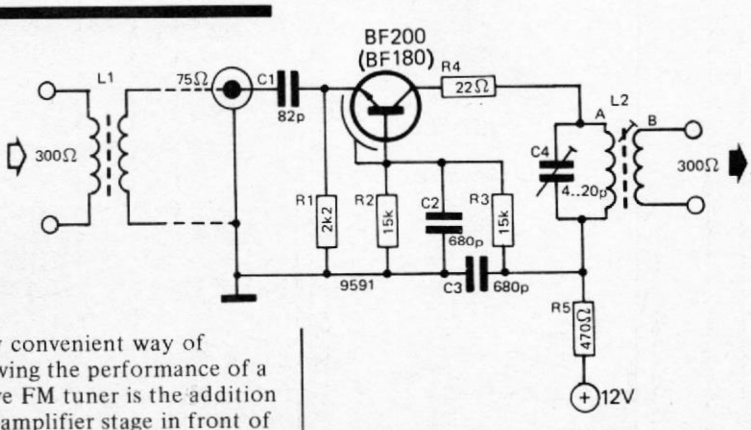




25 A very convenient way of improving the performance of a not-so-sensitive FM tuner is the addition of a VHF pre-amplifier stage in front of the existing equipment. Satisfactory gain improvement can be obtained if the booster stage is designed using a transistor having good VHF properties, such as the BF 200 or BF 180. These transistors also have good noise figures.

This booster design was tried in front of a variety of receivers with good results. The average voltage gain figure proved to be about 12 dB.

The preamplifier design is quite straightforward, using a common base configuration conditioned for minimum noise. 75 ohm antenna systems can be coupled straight to C1. For 300 ohm systems a matching transformer L1 can be made, by winding a 2 turn primary and a 1 turn secondary on a ferrite bead. Use 0.3 mm enamelled wire (SWG33). The output matching transformer is wound on a 6 mm (¼") diameter coil form. It should have a ferrite slug with a permeability $\mu r = 12$. Using 0.9 mm enamelled wire (SWG21) the primary (L2A) has 4 turns while the secondary (L2B) has 2 turns.

The circuit is not particularly critical to operate, provided that due care is taken in the construction of the amplifier and long connections are avoided. The transistor screening pin can be connected to the base, as shown in the drawing, or straight to the p.c. board common.

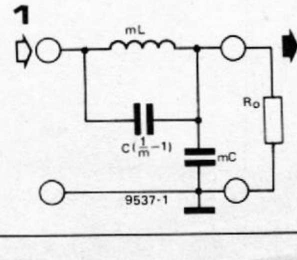
The only circuit requiring adjustment is the primary of the output transformer, L2/C4. This adjustment is carried out by tuning the FM set to a station at approximately 95 MHz. C4 is then set to minimum capacitance, after which L2 is set for maximum signal strength by adjusting the slug. Tuning is completed by the adjustment of C4, also for maximum signal strength.

26 Most RC noise filters show a fixed roll-off (slope) at frequencies above the cross-over frequency. Such filters with a single RC network usually have a 6 dB per octave roll-off.

Admittedly, with such filters, noise and high frequency distortion is made less obtrusive. However, not only the noise is effected, an important part of the high frequency content of the original signal may also have been wiped out. In such cases a filter with adjustable roll-off slope would be an asset.

Figure 2 shows a noise filter circuit with a cross-over frequency of approximately 7 kHz and a variable slope adjustable between 0 and approximately 25 dB/

1



$f_0 = \frac{1}{2\pi\sqrt{LC}}$

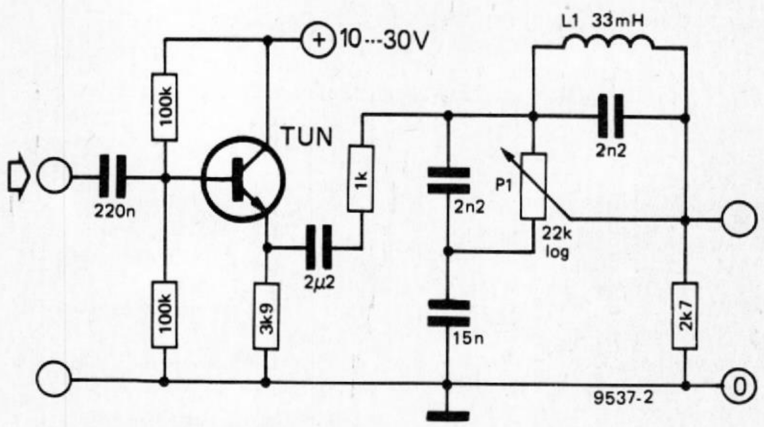
$R_0 = \sqrt{\frac{L}{C}}$

$m = 0{,}85 \ldots 0{,}9$

[illegible] va[illegible]ble slope [illegible] (c[illegible]nt.)
[illegible]7 active oscilloscope probe
[illegible]8 TTL-insurance (Norbert Conrads)

oct. The network, analogous to the cross-over filter in the Quad 33 pre-amplifier, is based on the m-derived low-pass filter (figure 1). The inductor is a Toko 33 mH 5% coil.
The frequency dependent characteristics of this filter are shown in figure 3. Graph 1 represent a 0 dB/oct roll-off over the audio frequency range; this is with P1 shorting out the inductor. Graph 2 shows a halfway position of the control, and graph 3 the effect of a maximum resistance across the inductor. The equation of figure 1 will be of guidance to the constructor who wants to design a filter after his own taste using other inductances and/or cross-over frequencies.

2



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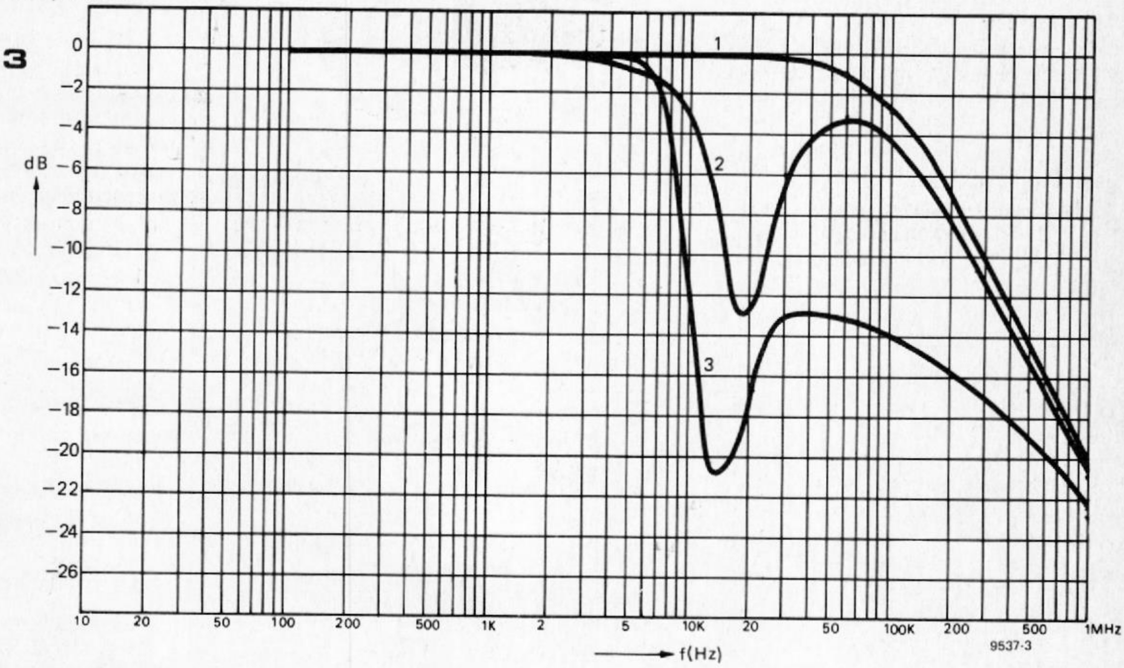
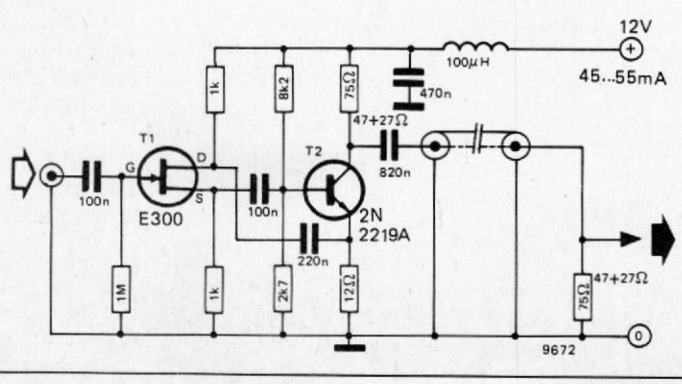


[illegible]7

The vast majority of oscilloscopes possessed by home constructors have a fairly low bandwidth, between 3 and 10 MHz. On the most sensitive voltage ranges the bandwidth is often further reduced. The usual piece of screened lead (coax) with prods has a lot af capacitance. This is particularly noticeable when testing logic or other pulse circuits, as this load capacitance can cause ringing.
An active scope probe acts as an impedance converter. It presents a high input impedance to the circuit under test and has a low output impedance to drive the cable capacitance. Gain may also be incorporated into the probe, which can make up for lack of gain in the scope.
A simple probe is shown in the accompanying circuit. A FET connected in source follower mode presents a high input impedance. This drives a bipolar transistor voltage amplifier. The upper frequency limit is approximately 30 MHz, provided sufficient care is taken with the construction. For high frequencies the lead must be correctly terminated at each end to minimise reflections.



28

It still happens too often that a mains supply fails due to thermal overload. The result is usually a short circuit between collector and emitter of the output transistor. When this happens in a TTL-supply, the output voltage may reach an extremely high value, usually causing destruction of the TTL-ICs. The circuit described below prevents such disasters because it switches off the supply if the output voltage goes higher than 6 V.