CQ·TV 81

THE BRITISH AMATEUR TELEVISION CLUB.

FEB 1973

THE BRITISH AMATEUR TELEVISION CLUB

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Members wishing to have material published in C Q - T V should send the manuscript and drawings to the Editor; articles are invited on all subjects of interest to amateurs and should be of about 1500 words; larger articles should be divided into convenient Parts for publication in consecutive issues of the journal.

COVER PHOTO Oscar 6 SSTV Pictures

EDITORIAL

As a result of the last committee meeting there is a slight change to the list of Club Officials in that Gordon Sharpley takes over the post of Membership Secretary from Nick Salmon, who has stood down after a long and hard-working spell in the post. Malcolm Sparrow remains Chairman and sends this message to members:

Dear Members,

The first meeting of the new B.A.T.C. Committee elected at the September General Meeting of the Club was held at Wolverhampton on December 10th last.

At the meeting I was elected to serve as your Chairman for the next two years and I would like to draw your attention to the new list of Club Officers which appears in this issue of C Q - T V In addition to welcoming those new members to our committee I must also voice a sincere vote of thanks to those retiring committee members who have now stood down and thank them for their support and efforts during their term of office.

Looking forward, the future holds a challenge to us all with a smaller 70cm frequency allocation in the U.K. for amateur vision of 432 - 440MHz, which we in turn have to share with the sound amateurs in 432 - 433.5MHz and the amateur satellite signals around 435MHz.

On the brighter side by negotiation with the R.S.G.B. the 70cm beacon stations are shortly to be moved to around 432MHz to minimise interference with amateur tv. In all we should perhaps look to our own equipment and try to improve our atv reception techniques - why use a wide open front end tuner and get co-channel interference? In all a narrower atv band on 70cm should lead to improvements in operating techniques too. With more stations becoming active and atv contests being organised, it is up to us to match the problems with new methods.

As amateurs we experiment in the art of communication and yet, being honest with ourselves, we do not seem to communicate with each other very well. For a start, make a New Resolution for 1973 to let others know of your progress and problems. No doubt others have had the same problem and may have found the answer. To this end use C Q - T V. Air your views and news - it is your magazine so you should use it.

We hold a General Meeting of the Club every other year and each time we usually try to suit the majority of our members by holding it in London. I know that for some of our British members this is too far to travel, so why not organise your own mini-Convention.

continued on page 16

CONTENTS

Committee Member's Addresses	page 1
Editorial	page 2
Circuit Notebook No. 12	page 3
European Reporting System	page 6
Ideas for Amateur Colour Part 5	page 7
1972 atv Contest Results	page 10
Receiving Amateur tv	page 11
Slow Scan News	page 14
A Slow Scan Scanner	page 17
Integrated Circuits Part 11	page 21
Postbag	page 31
Adverts	page 32

CIRCUIT NOTEBOOK No 12

J. Lawrence GW6JGA'T

Camera Tube Scanning Circuits

The primary requirement of the line or field scan deflection circuit is that it should provide a linear scan and rapid retrace or flyback.

As the scanning is done magnetically, this means a linearly changing current during the scan and a rapidly changing current during the flyback.

Transistor Line Scan Circuits

Most simple transistor line scan circuits rely on the linear rise of current that occur in an inductance (the line scan coils) when a voltage step is applied across it.

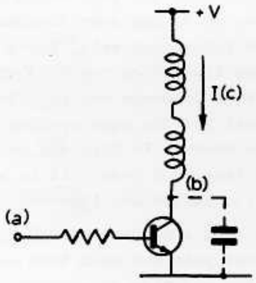


Fig. 1a

The basic circuit is shown in Fig. 1a. The transistor is used as a switch with the line scan coils in the collector circuit.

To start with, let us assume that the base of the transistor is reverse biassed and that the collector is at the supply voltage. At the start of the line scan the transistor is switched on hard by a suitably

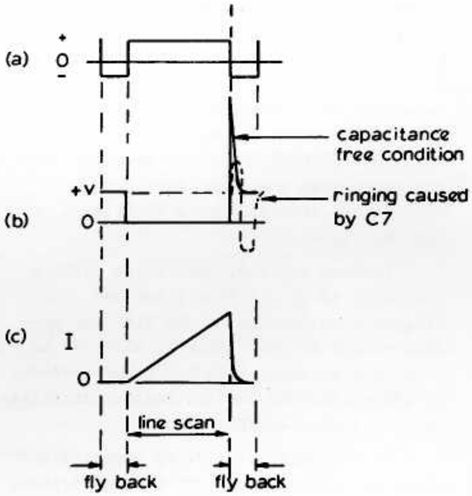
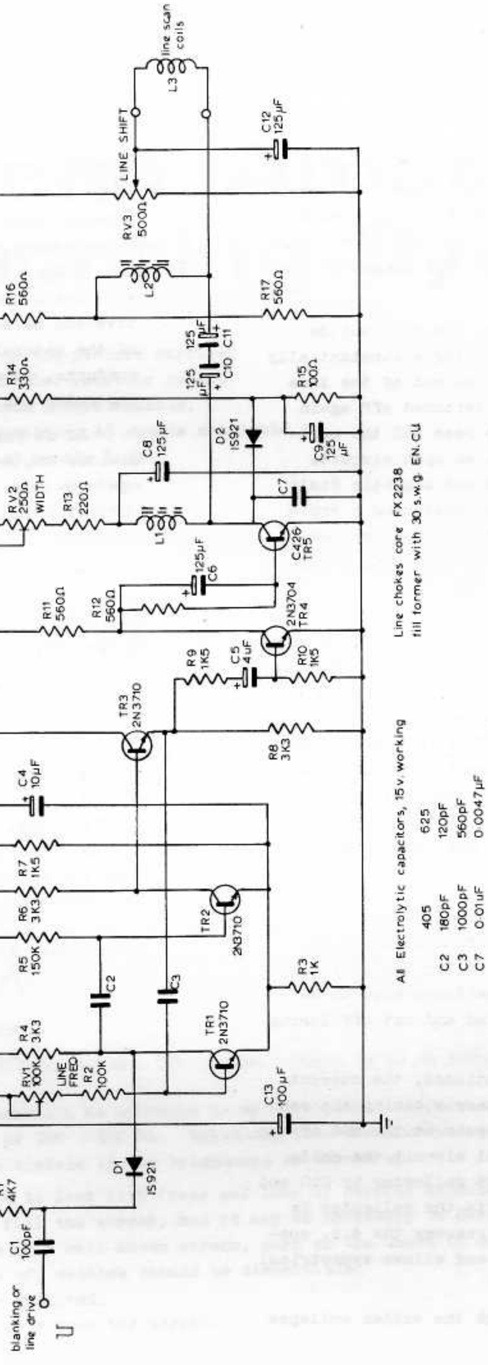


Fig. 1b



blanking or line drive
+12v
C1 100pF
R1 4K7
D1 IS921
RV1 100K LINE FREQ
R2 100K
R4 3K3
TR1 2N3710
C13 100µF
C2
C3
R5 150K
R3 1K
R6 3K3
TR2 2N3710
R7 1K5
C4 10µF
TR3 2N3710
R8 3K3
R9 1K5
C5 4µF
R10 1K5
R11 560Ω
R12 560Ω
TR4 2N3704
C6 125µF
TR5 C426
RV2 250Ω WIDTH
R13 220Ω
L1
C8 125µF
C7
D2 IS921
C9 125 µF
R14 330Ω
R15 100Ω
C10 125 µF
C11 125 µF
R16 560Ω
L2
R17 560Ω
RV3 500Ω
LINE SHIFT
C12 125µF
L3 line scan coils
Line chokes core FX2238 fill former with 30 s.w.g. EN.CU
All Electrolytic capacitors, 15v. working
405 625
C2 180pF 120pF
C3 1000pF 560pF
C7 0.01uF 0.0047µF

Fig. 2

large positive current applied to the base. The collector falls immediately to almost zero volts and the scan coil has a voltage step almost equal to the supply voltage across it.

The current through the coil builds up in a linear manner giving a substantially linear deflection. At the end of the line scan the transistor is switched off again by reverse biassing the base and the collector once more becomes an open circuit. This causes the current and magnetic field in the coils to collapse producing a rapid deflection in the opposite direction and generating a reverse voltage having a high peak value as shown in Fig. 1b.

A complete practical circuit of a line scan generator is shown in Fig. 2. In this circuit the line oscillator consists of Tr1, Tr2 and Tr3 connected to form an astable multivibrator. This is synchronised at line frequency by negative going line drive or blanking pulses applied to C1. Emitter follower Tr3 is included to ensure good waveshape at C5. The line frequency is set by RV1.

The driver stage Tr4 is turned off during the line period and on during the flyback, this results in Tr5 the line output transistor being switched hard on during the line scan period and cut off during the flyback.

As previously mentioned, the current in the coils rises linearly during the scan until Tr5 is cut off again at the end of the line. In the practical circuit the coils are a.c. coupled to Tr5 collector by C10 and C11 and the d.c. feed to the collector is via a choke L1. This removes the d.c. component from the coils and allows symmetrical deflection.

During the flyback the sudden collapse of current causes the scanning coils, in parallel with C7 the flyback capacitor, to oscillate at their natural resonant frequency, shown dotted in Fig. 1b. However, D2 will not allow the collector to swing negative and as soon as the negative half-cycle of the natural oscillation commences D2 conducts, driving current through the scanning coils and starting off the next scan.

D2 is forward biassed by the potential divider R14 and R15 to ensure a smooth change over from reclaimed scanning current to current provided by the line output transistor Tr5.

An additional choke L2 is provided to enable shift currents to be passed through the scan coils for picture centring. The SHIFT control is RV3.

Scan amplitude is determined by the voltage applied to the output stage and this is adjustable by RV2 the WIDTH control.

With a 12Volt supply rail and normal direct drive transistor type scan coils, this circuit, which is based on the Link 101 Camera, will provide more than adequate scan for a Vidicon tube with integral or separate mesh.

References Millman and Taub
"Pulse Digital and Switching Waveforms" Page 572
McGraw Hill

Instruction Manual for the EMI type 8 Camera

Instruction Manual for the Link 101 Camera

European Amateur tv Reporting System

Picture Carrier

B0 Nothing receivable from the picture carrier.
B1 A3 sound or speech audible, receiver on AM.
B2 A3 sound visible, speech understandable.
B3 Non-lockable picture visible, A5 rumble available.
B4 Lines can be locked, A5 rumble loud.
B5 Lines and picture can be locked.
B6 Call readable.
B7 Persons recognisable.
B8 Details recognisable.
B9 Picture almost free from noise.
B9+ Picture completely free from noise.

N.B. The tv receiver is switched to AM for B1 - B4

Sound Carrier

T0 Nothing receivable from the sound carrier.
T1 Test tones audible, speech unintelligible.
T2 Speech sometimes understandable.
T3 Speech understandable when picture is at black level.
T4 Speech understandable when picture contest is white.
T5 Speech understandable if tuned for best sound.
T6 Poor speech understanding if tuned for best picture.
T7 Good speech understanding if tuned for best picture.
T8 Sound almost free of distortions if tuned for best picture.
T9 Sound completely free of distortions if tuned for best picture.

Remarks

B1 The tv receiver is switched to AM. The vision carrier is to be modulated by speech in the A3 mode.
B2 Again the tv receiver should be switched to AM. Horizontal bars should appear on the screen if the sound is tone of 200 - 800 Hz. Whistling into the microphone is an alternative.
B3 Sync pulses should be visible if the brightness is turned up.
B4,5 It should be possible to lock both frame and line by careful adjustment.
B6 The call sign should fill the screen, and it may be necessary to darken the shack to read it.
B7 The picture should be of a well known person, such as the Queen or some local personality.
B8 Scales, and the hands of, watches should be discernible.
B9 3MHz on 625 should be resolved.
B9+ 400 v should be available from the aerial.

Ideas for Amateur Colour
Part 5

Nigel Walker
G6ADK'T

The amateur who has managed to produce colour captions by simply feeding the output of a monochrome camera into the R,G or B input of a coder or a monitor will feel that the next step should be to build a proper "synthesiser", where the artificially produced colours are infinitely variable.

The synthesiser to be described splits the video input (from the monochrome source) into three levels, such that a different colour can be set up for each level. A total of nine potentiometers are thus required to control the colour output.

Figure 1 shows the actual level splitter. The video is first clamped, and then fed into two level detectors, the switching level of which is made variable by means of preset potentiometers. The outputs from the two level detectors are then fed to a gating system; for output "A" (highlights) the signal is inhibited by blanking, but is otherwise unaffected. Output "B" (lowlights) is inhibited by blanking too, but it is also inhibited by "A" thus producing a signal corresponding only to lowlights, and not lowlights plus highlights as would otherwise be the case. Output "C", also inhibited by blanking, produces an output when "A" or "B" are in a low state and thus produces a signal corresponding to the blacks, or background of the picture.

Figure 2 shows the output stages. The nine potentiometers provide independant adjustment of the proportions of red, green or blue for each of the three levels, and are the controls which should be mounted on a front panel. The output filters are provided to limit the frequency spectrum of the outputs to that corresponding to the 625 line system bandwidth.

With careful use of crayons applied to a caption made with white "Letraset" to produce three brightness levels (i.e. black, grey and white) quite reasonable three colour captions can be produced using this synthesiser. Very colourful effects can also be produced by pointing the camera at an ordinary scene, and there is much scope for artistic experiments along these lines.

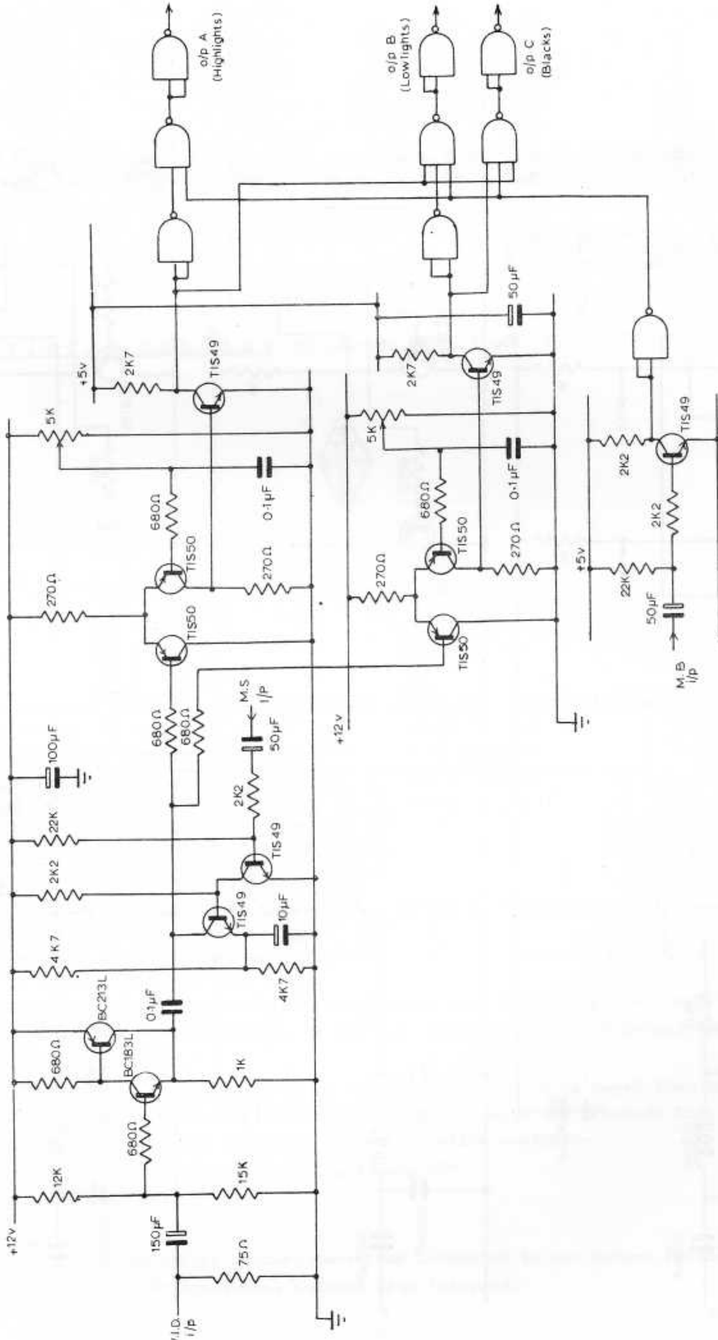


Fig 1.3 Level splitter

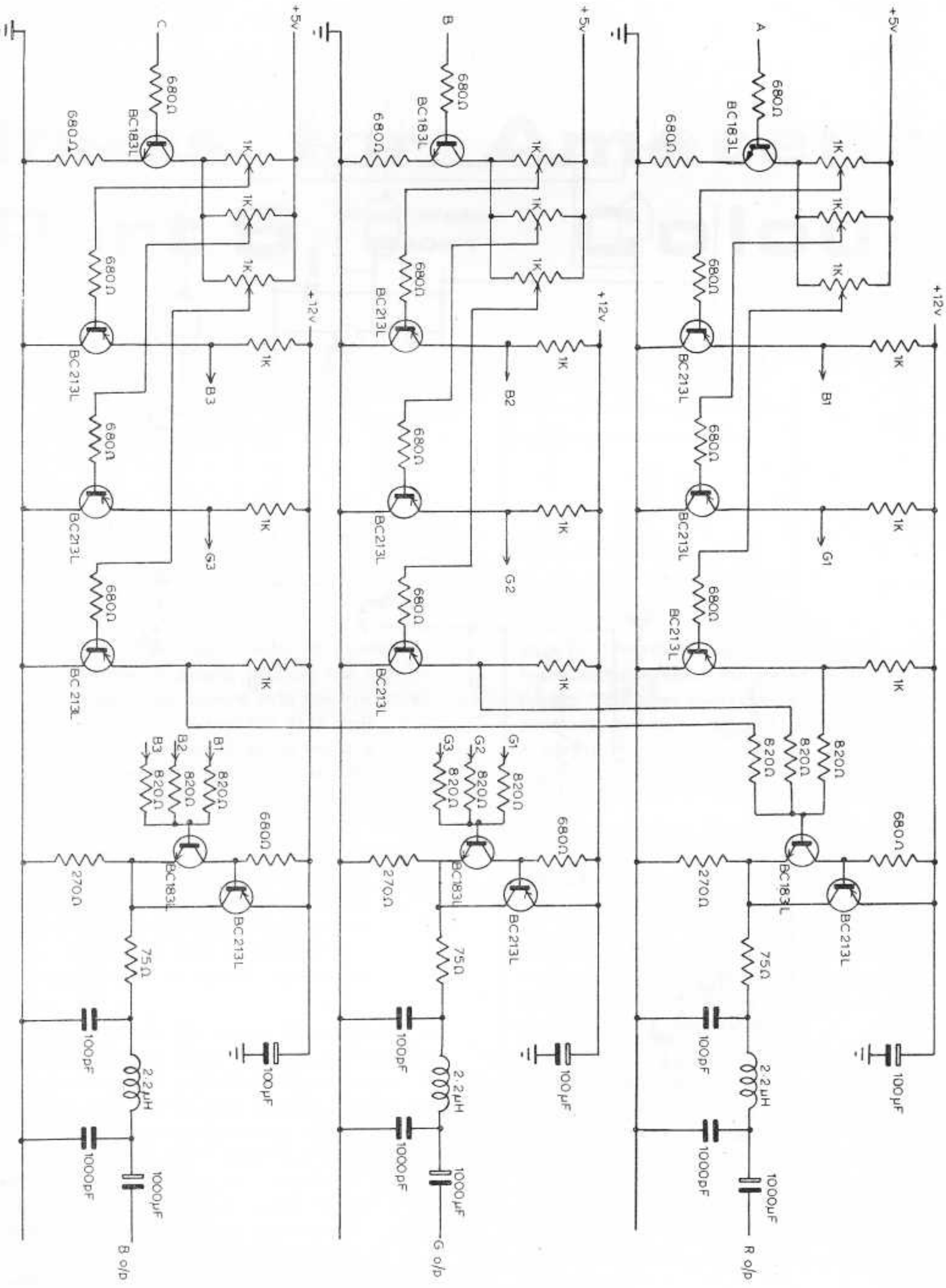


Fig 2. OUTPUT STAGES

1972 atv contest results.

B.A.T.C. N A T I O N A L C O N T E S T 23 -24 -30 SEPT, 1 OCT. 1972

SECTION ONE

PLACE	CALL	POINTS	No. of CONTACTS	BEST DX	No. STATIONS CALLED/WORKED
1	G6AEV/T Portable	4294	58	G6AFW/T 165Km	12 - 8
2	G6AGT/T/A	3309	54	G6AEV/T 135Km	5 - 18
3	G6AFW/T	3102	31	G8ARM 225Km	11 - 8
4	G6KQJ/T	1450	18	G8ARM 203Km	6 - 9
5	G6GDR/T	679	13	G6OPB/T 46Km	4 - 3
SECTION TWO					
1	G3YQC	712	6	G6AGT/T 68Km	2 - 0

Some twenty one /T stations took part altogether, with thirty one sound only stations. Conditions for the first weekend were above average, which added to the normal range expected for atv.

It is a pity that so few stations submitted an entry, and it is hoped that more will do so with the atv cumulative contest now being held. Please note one omission from the published rules, in that there will be two sections to the cumulative contest:-

A. For those stations transmitting atv

B. For all others.

The results of the September contest have been forwarded to our German friends, and we await the results of the International Contest with interest.

Receiving Amateur tv for the Beginner

Using the Mullard ECL1043 Varicap Diode Tuner.

by Malcolm Sparrow

To receive Amateur Television these days need not be as hard as it used to be in the very early days when A.T.V. had just started; in fact, using the new Mullard ECL 1043 tuner unit makes it just about as easy as possible.

The one saying to remember when trying to receive any UHF signals is that "A chain is only as strong as the weakest link in it". In other words, provided that there is an ATV signal on the air, to receive it one requires: A 70cm aerial, a coaxial cable to connect it to the tuner unit, a tuner unit, and a tv receiver. If any of the items does not function properly then the results will be impaired.

FIRSTLY THE 70CM AERIAL

These days it is rarely worth trying to design and construct your own aerial. J. Beam Engineering Ltd. of Northampton produce a range of 70cm aerials and the beginner would be well advised to start off by purchasing one of their aerials such as the 18 element parabeam (Cat. No. 70/18P,) price £7.45p, which has a gain of 17dB over a plain dipole. (Do not use their 14 element skybeam aerial for ATV as this aerial does not have a suitable band width for ATV.

Whichever aerial you finally decide upon, the next thing to consider is where to put it. In practice it will pay to try to mount the aerial so that it is clear of all the roof-top obstructions as wet tiles can make quite considerable signal attenuation at 70cm. Also some provision for turning the aerial should be made to enable an incomming signal to be peaked for maximum. This may not coincide with the most direct route to the transmission source.

NEXT, THE AERIAL COAXIAL DOWNLEAD

Having decided upon which aerial to use and where and how you are going to mount it the next thing to consider is the coaxial downlead.

This should be as good a cable as you can obtain and afford. There is no point in erecting a high gain aerial only to loose all the signal before it gets to the tuner unit. One should use a good quality low loss BBC 2 type coaxial cable such as "Aerialite Type M4205 Aeraxial Cable". This is a ⅜ inch diameter coaxial with a cartwheel cross-section. That is, the insulator between the inner and outer cables is partially air-spaced. A point often overlooked when using this type of semi-airspaced coaxial cable is that although the air in these holes in the cable is dry when you purchase the cable new, it will soon become damp when connected to the aerial way up in the air.

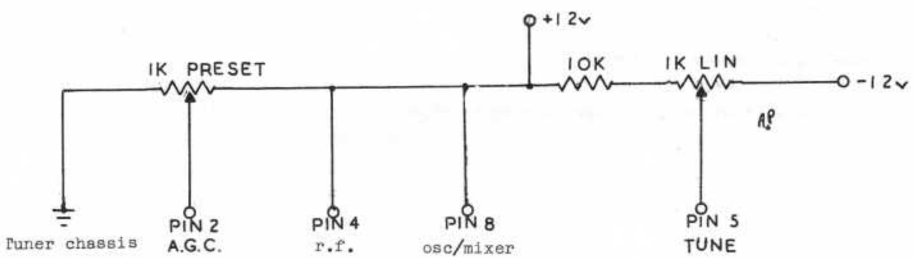
Whilst the cable will give first-class results when.first connected to the aerial it will soon deteriorate in performance if steps are not taken to trap the dry air inside the cable. This is simply acheived by cutting the cable to the required length to be used - always keep it as short as practical and do not leave large lengths in use just to avoid cutting a run of coaxial cable - and then by melting the insulator at each end of the cable, seal each end to keep the dry air in the cable before attaching it to the aerial. This can usually be done quite easily by using a dry soldering iron, but take care not to let the inner and outer cables short out whilst doing this.

NOW TO THE TUNER UNIT

The Mullard Varicap Diode Tuner Unit type ELC 1043 is available from Manor Supplies Ltd., at £4.50p each plus 25p post & packing. This tuner is built in a tinplate box with detachable sides and is tuned by varying the voltage applied to the Varicap diode in it; it should be connected up as follows:

The aerial lead should be connected to the input terminal which is on one end of the unit, the outer of the coaxial cable being soldered to the tuner chassis.

Next, refer to the pins and holes which are located in the base of the tuner unit between the four mounting lugs. Commencing from the end of the tuner which has the aerial input terminal on it number these holes (some of which have pins in them) 1 to 11 and connect them up as follows:



Transistor Supplies.

The 1Kohm linear Variable potentiometer feeding pin 5 is the tuning control, whilst the 1K preset control should be set to feed pin 2 with 3v +ve to the tuner chassis. The output from the tuner is taken from pin 10 with the braid of the output coaxial cable being connected to the tuner chassis.

The output lead should be taken to the aerial input socket of the tv receiver and the receiver set to channel one. With the tv set turned on and 12v dc. applied to the tuner (and the aerial connected to the tuner), you are now all set to watch for ATV pictures provided that there are some on the air to start with, of course.

Once a signal has been received the I.F. output coil slug in the tuner can be tuned for the best signal but there will probably be only a slight improvement.

This equipment will provide an up-to-date ATV receiving system which will be on a par with any valve tuner plus low noise transistor preamplifier, as the Mullard Tuner noise factor is quoted as being 8dB.

It will of course be necessary to change over the detector diode polarity in the tv set and also the link time base frequency if you wish to watch 625 line -ve modulation ATV as the system as described is intended for 405 line +ve modulation ATV as mainly used by amatuers outside the London area.

Finally, do not put more plugs and sockets in your 70cm aerial down lead than absolutely necessary, as each time you use one you will lose a little of the signal.

NOTE: If you want to tune the whole of the television band IV with your tuner it will be necessary to arrange for the tuning voltage on pin 5 of the tuner to be varied between 0.3v and 28v +ve, but the circuit as shown should tune the amateur 70cm band without any need for internal modification to the Mullard Tuner unit.

References: J. Beam Engineering Ltd.,
Rothersthorpe Crescent,
Northampton.

Manor Supplies Ltd.,
172 West End Lane, London, N.W. 6.
Mail order 64 Golders Manor Drive, London, N.W. 11.

Data Sheet on UHF television tuner type ELC 1043
Mullard Ltd.,
Mullard House,
Torrington Place,
London, WC1E 7HD

THE C Q - T V S.P.G.

At the time of going to press 99 spg boards have been sold to B.A.T.C. members - together with 47 Genlock boards. Who is going to be the lucky fellow to get number 100? This represents about 10% of B.A.T.C. membership.

In reply to a mainslock modification as suggested by david Wilkinson on page 32 of this issue; this has been tried and found to work. It is not the complete answer though. The 50Hz phase modulation is reduced from about 11us to about 8us but the pull in speed and locking power are improved too. One slight problem is that it seems to make the verticals rather kinked. There is no effect on the Genlocking performance.

"SLOW SCAN TELEVISION"

Some members who have bought this booklet unfortunately did mot receive an errata slip. Please note that in Figure 10 on page 6, the base of the BFX29 should be connected to the collector of the BFX85 and NOT to the emitter.

Slow Scan News

Dear Fellow Slow Scanner,

As you probably know Oscar 6 is now in orbit and several of us have had success in sending slow scan television through the satellite. On early evening passes we occasionally hear "G" signals on sideband and CW. We would very much like to work a slow scan contact with a DX station.

If you are interested please drop me a line immediately. I suggest if you are interested that you sked me on 11.30GMT on Wednesdays on 14230Kc/s and we can arrange for a contact through the satellite. Oscar 6 is turned on on Thursday, Fridays, Saturday and Sunday evenings as well as Saturday and Sunday mornings. This saves on battery power since one of the solar panels is having trouble. If you skeded me on Wednesday we would be ready for weekend passes. Another way to alert us is to contact the slow scan television net which meets at 1800GMT on 14230Kc/s each Saturday afternoon. I am one of the net controllers and can always be reached directly or by relay from the Atlantic coast at that time.

The procedure for making contacts is to broadcast continually on your chosen frequency for the full 20 minutes of the pass and to check later to see if you were seen or heard. The early pass of the evening is the only one of that group where we have a chance of seeing each other across the Atlantic. WA9UHV and myself are operating on 145.930 and 145.940 respectively. This brings us out at 29.480 and 29.490MHz respectively. If you can sked me I can alert you to the possible passes where we might have success.

Please let me hear from you. Satellite communications are a great deal of fun and are very exciting.

Don C. Miller,
Waldron, Indiana,
Box 95, 46182,
U.S.A.

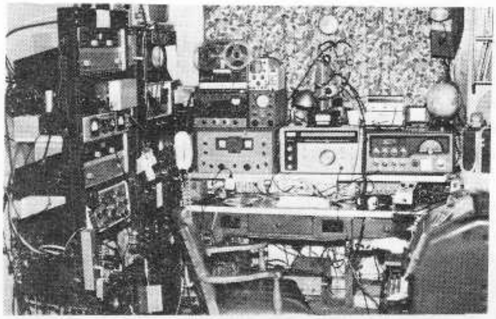
OSCAR 6

Oscar 6 was launched on 15th October 1972 from Vandenburg, California, piggy-back fashion with a weather satellite. The Oscar translator was designed and built by the AMSAT organisation, a group of radio amateurs dedicated to forwarding the technology of Radio Communications.

The translator has an input band around 145.5 and an output band from 29.45 to 29.55MHz. All modes of radio communication can be translated.

The orbit of the satellite is polar with an inclination of 101.73 degrees and a period of 114.5 minutes. There is a telemetry beacon on 29.45MHz. The "CODESTORE" memory of Oscar 6 is capable of being loaded with ground based information while in orbit. W1AW, the official

American Radio Relay League station, constantly transmits undating information to all interested radio amateurs on the QST published frequencies. Announcements are made in both 'phone and RTTY.



The photos show W9NTPs gear which made the first two way QSO through Oscar 6. The first contact was made on orbits 30 and 41 and by October 40 pictures had been exchanged with WA9UHV. The gear at W9NTP comprises a Heath SB101 high frequency transceiver with 30w out, a homebrew linear 100w 2m amplifier with a 4CX250 in the final, a 10 element vertically polarised 2m antenna, rotable in azimuth only, a 10m 4 element rotable Yagi, a NX-303 National Hamband receiver on 10m and a homebrew sampling SSTV camera with 7 inch monitor. Similar equipment is used at WA9UHV except that the aerials are rotable in azimuth and elevation.

'9NTP reports that QSB is so severe that it is difficult to get even one complete SSTV frame. The main problem seems to be "power robbing" by other stations, but as can be seen by the photos (see this page and the front cover), some success has been possible.

ROBOT EQUIPMENT (especially Camera Model 80)

Richard Thurlow, G3WW, of Wimblington, Cambridgshire has sent us some information about a new specification for SSTV, as used by the M.P.T., resulting from some considerable correspondence over the last few months. This is how G3WW tells the story.

"To those of you who I have pestered for information by phone calls, cablegram and letters as to why and how the ROBOT camera with its 128 lines is allowed to be used universally in both the 60Hz and 50Hz AC Mains areas of 120 lines standard, I hasten to report that after passing on to London the excellent information gathered, especially from K6IV himself, W4BW the Chief of the Amatuer and Citizens Division of the FCC. thro my good Attorney friend of 1960 Washington DC visit W3GZ/4, and K1PLP Asst. Tech Ed. QST who reviewed the ROBOT equipment, that in the USA the 120 line standard is one adapted by the US Amateurs and is not regulated by the FCC, and that ROBOT have departed from the accepted standards in two respects on the grounds that the performance is improved and the signal from the camera is received well by all monitors i.e. 128 lines per picture and Vertical sync pulse duration 66 milliseconds rather than 30 milliseconds, I have today received permission from the Ministry of Posts and Telecommunications (per Mrs. A.I. Campbell) to use SSTV under fresh conditions.

'our engineers have now reviewed the technological aspects of the equipment required to meet our specifications and I have pleasure in advising you that we are prepared to grant you permission to use the following bands: 7-7.10MHz, 14-14.35MHz, 21-21.45MHz, 28-29.7MHz, 144-146.0MHz two year period etc. P.S. Enclosed is a copy of our revised specification which covers ROBOT 80'.

SLOW SCAN TELEVISION SPECIFICATION

Number of lines per frame 128 ± 8 lines
Aspect ratio 1:1
Horizontal frequency (frame) $16\frac{2}{3} \pm 1$Hz
Vertical frequency 7.68 secs. (limits 6.79 to 8.68 secs.)
Horizontal sync pulse 5 milliseconds
Vertical sync pulse 30 milliseconds (nominally)
F.M. subcarrier sync 1200Hz
black 1500Hz
white 2300Hz

Those of you who were NOT pestered might like to know that the necessity for all the above arose from the fact that when applying to transmit SSTV I (and as later discovered G3NMH, also) sent up with my application ROBOT's printed brochure packed with their Model 70 Monitor WHICH SETS OUT THEIR SPECIFICATIONS FOR USE WITH USA 60Hz AC Mains ONLY; permission could not be granted ON THESE FIGURES; ability to perform on 50Hz AC Mains quickly explained by provision of 'a changeable jumper' in the ROBOT Camera, but there remained the 128 lines against the world wide adopted 120 lines.

Get some ideas from other members in your area (the Treasurer will tell you who they are) and we will try to get at least one Committee member to attend to bring you up to date. Just give us a reasonable amount of notice; remember it takes approximately ten weeks to get a letter into print in C Q - T V.

Finally may I wish you all you wish yourselves for 1973 and here's to a more successful, and dare I say, colourful atv year to all our members and <u>let us all remember to try to communicate.</u>
Malcolm Sparrow G6KQJ/T
Hon. Chairman B.A.T.C.

Once again subscriptions are due - it is only one pound, after all. And those who don't pay, won't receive C Q - T V. Please send to the Hon Treasurer, whose address is on page 1, and this time we have a little extra to ask of you. To enable us to bring our records up to date, if you have a Postcode, could you let us know what it is. Also, if the address on the envelope you received this C Q - T V in is wrong in any way, please let us know about that too. This will help us, the Post Office, and ultimately, yourselves.

At the next vhf Convention, to be held as usual at Whitton, B.A.T.C. would like to have an organised display, if at all possible. Dave Lawton G6ABE/T has taken on the responsibility for coordinating the effort, so if you will be there and would like to help, or have some equipment you would like to show, please contact Dave - his address is printed on page 1.

THE EDITOR.

A Flying Spot SSTV Scanner

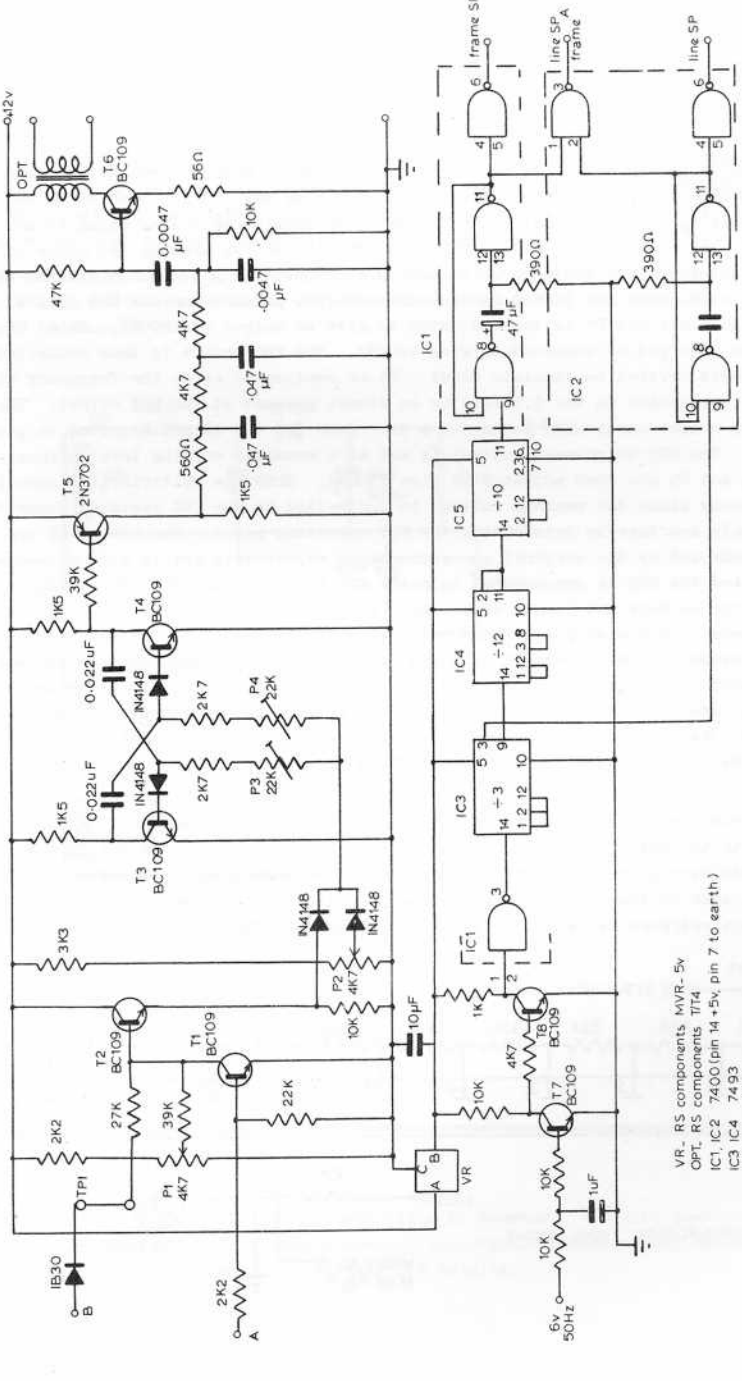
B. J. Arnold M.A. G3RHI

This article concentrates on the electronics of the scanner rather than the optics. It has been used at G3RHI without an optical system to make tape recordings for transmission by placing cutouts directly on the face of the station SSTV monitor tube. The contrast control on the monitor is turned right down and the monitor raster is synchronised from the sync pulse generator in the scanner. Thin card cut-outs or transparent acetate with black letraset letters are placed firmly on the tube and the 931A is stood about 12 inches away. Care is taken to minimise light leakage between the card or acetate and tube but provided this care is taken good black and white captions and simple cartoons are recorded. An optical system using 35mm transparencies would be a real improvement but so far the search for readily available parts for an easily built lens and transparency changeover system has not been successful. By contrast all the parts for the electronics are readily obtained.

The power supply is conventional. The Crofton Electronics transformer, though not essential, might well have been designed for the job. The 12 volt supply is standard. The 5 volt regulator was placed on the modulator printed circuit board and wired in permanently to minimise the risk of connecting 12 volts to the I.C.s. The photomultiplier EHT is a standard full wave voltage doubler connected to the 200+ volt HT winding on the Crofton transformer. The load is a modest 1mA but safety precautions should be taken since with this transformer the EHT voltage is some 480 - 500 volts.

This is enough for the 931A. Higher EHT can be used so long as the total voltage is measured and the resistor connected between X and the EHT calculated so that the total current through the chain is 1mA using the values in Figure 2. With EHT connected the tube should not be fully exposed to daylight or bright light. One of the old 1¼ inch coil forms or a piece of plastic tube with 3 strips of foam draught excluder inside make a good shroud. The output voltage of the tube is negative and a positive voltage is needed to drive the modulator. The 2N3819 buffer also conveniently inverts the signal polarity and its 2Mohm input control should be readily adjustable. Screened cable is used between the FET and point B on the modulator.

The modulator circuit includes a sync pulse generator.T7and T8 square the 50Hz voltage which is divided in IC3 by 3 and shaped by IC2 to provide line sync. IC4 and 5 together divide by 120 to give frame sync shaped by IC1. The heart of the modulator is the multivibrator around T3 and T4. A ganged pot is not available for P3 and P4 but the adjustment of separate



VR. RS components MVR-5v
OPT. RS components T/T4
IC1, IC2 7400 (pin 14 +5v, pin 7 to earth)
IC3 IC4 7493
IC5 7490

Fig. 1. SSTV Modulator and sync. pulse generator

pots in step is easily done. T5 is a buffer and a square wave is present at the collector which is followed by a low pass filter. The result is a sine wave at the base of T6 which is an amplifier to make up for the losses in the filter. On receipt of a sync pulse T3 and T4 must produce 1200Hz, and black 1500Hz and on white 2300Hz. T1 and T2 provide the means of varying voltages to control this frequency. To set up the modulator a frequency counter is essential. For initial adjustment the 2N3819 is disconnected from point B and the SPG is disconnected from point A. The black pot P1 is then adjusted to give an output of 1500Hz. Point TP is then earthed and the sync pot P2 adjusted to give 1200Hz. The FET output is then connected to point B with the 931A covered to simulate black. P1 is readjusted since the frequency will rise a few cycles in response to the 0.5 volt or so always present at the FET output. The tube is then exposed to a bright continuous raster in a dark room and the 2M pot adjusted to give 10 volts at point B. The CRT brightness control is set at a sensible working level without any colour filter. P3 and P4 are then adjusted to give 2300Hz. Thus the multivibrator cannot go above this frequency since the maximum voltage is controlled by the 10K resistor connected to the 12 volt supply and thereby determining the FET operating point. The black and sync frequencies are then rechecked by the original procedure since adjustments are to some extend interdependant. When completed the SPG is reconnected to point A. Prior to this the SPG outputs can be checked on a scope for voltage and length of pulses. Even with the so-called new but cheap I.C.s there should be some 2 volts at point A and only 0.6 volts is needed at the base of T1. T1 is a transistor switch which on receipt of a sync pulse bottoms its collector to achieve the same effect as earthing TP during the setting up procedure. Throughout all adjustments a small speaker connected to the transformer wads it and gives a good indication of success or lack of it. The unit is now ready for operation. The Mark 1 version was built on an open chassis while Mark 2 was built in a diecast box with a rectangle cutout to expose the sensitive area of the 931A.

Lucky the operator who has two monitors or a monitor and a separate CRT unit producing a raster. With the raster set up in the dark about 12 inches away from the 931A and synchronised from the modulator a facsimile of the card cut-out or transparency placed on the face of the tube can be seen on the monitor to which the modulator is connected. If a second monitor is not available patience is necessary while the picture is tape recorded and then played back on

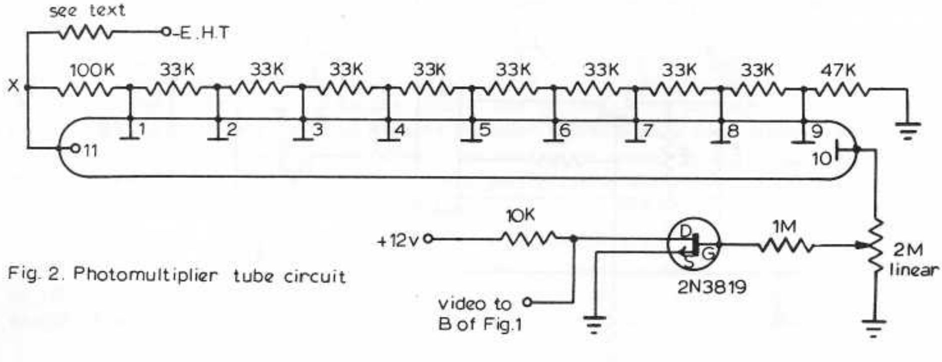
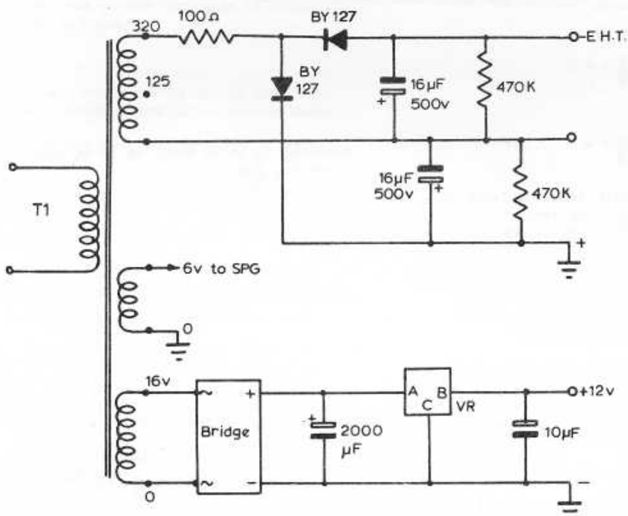


Fig. 2. Photomultiplier tube circuit

the monitor originally used to provide the raster. Although time consuming, with experience results are good since the only two variables are the CRT brightness control and FET input. In practice the FET is set and the brightness is varied. Operation is best done in the dark so tape recording rather than direct transmission is more convenient.

FOOTNOTE: The author has a few printed circuit boards for the combined SPG and modulator priced £1.60. P1,2,3 and 4 are available fitted to this board at 40p the set. The 931A and base are stocked be Henry's Radio.



VR_ RS components MVR 12

Bridge_ RS components REC 60

T1_ available from Crofton Electronics_ same as Camera Kit Transformer

Fig. 3. Power supply

NEWSFLASH

A TV Convention is to be organised in 1973 to take place in Germany. B.A.T.C. members who are interested in attending this Convention should contact the organiser Manfred May DC6EU, Caesarstrasse 13, Bayenthal 51, D5000 Koln, Germany for details.

INTEGRATED CIRCUITS

PART 11

A.CRITCHLEY Dip El; C Eng; MIERE.

MORE ON OPERATIONAL AMPLIFIERS

FREQUENCY COMPENSATION

This is a complex subject which tends to frighten people off Op. Amps. You may feel the same after this description - in some detail to show the reasons for the various methods of compensation.

The basic reason for frequency compensation of high-gain, high-frequency amplifiers is to prevent instability. It consists of adding phase-changing networks to the amplifier to lower its gain at frequencies where the phase after feedback would cause oscillation.

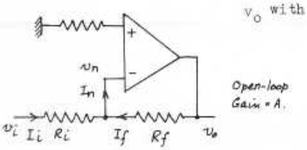
Phase-shifting of the high-frequency response of a high-gain amplifier is inevitable in circuits containing many transistors - as Op. Amps. do. The technique used in taming these amplifiers is to ensure that the gain of the system at the troublesome phase is less than unity.

So what is involved?

The basic inverting amplifier uses a resistive feed-back network as shown below and has a gain of -Rf/Ri - assuming that the open-loop gain is infinite and that all other factors are perfect.

If, however, the open-loop gain, A, is not infinite as is the case for a practical amplifier, and especially at high frequencies, then the expression is no longer true. Normally the slight voltage, v_n, at the input terminal is ignored but now it has to be taken into account.

v_n is required to generate v_o with a gain of A



For an inverting amplifier the sum of the currents at the inverting input is zero. i.e. $I_i + I_n + I_f = 0$.

However, I_n is so small as to be negligible in any but the worst Op. Amps.

Thus $I_i + I_f = 0$ or $I_i = -I_f$

Also, $v_o = -v_n.A$ or $v_n = -v_o/A$

For a normal low-frequency amplifier it is the latter term which is ignored. Hence $G = v_o/v_i = -R_f/R_i$

For the finite open-loop gain case;

$$I_i = (v_i - v_n)/R_i \qquad \text{and } I_f = (v_o - v_n)/R_f$$

$$\text{So } \frac{v_i - v_n}{Ri} = -\frac{v_o - v_n}{Rf} = \frac{v_n - v_o}{Rf}$$

$$\text{or } v_i - v_n = (v_n - v_o).Ri/Rf$$

$$v_i = \frac{Ri}{Rf}.(v_n - v_o) + v_n$$

$$\text{but } v_n = -v_o/A, \quad v_i = \frac{Ri}{Rf}\left(-\frac{v_o}{A} - v_o\right) - \frac{v_o}{A}$$

$$= -\frac{Ri}{Rf}.v_o\left(\frac{1}{A} + 1\right) - \frac{v_o}{A}$$

$$= -v_o.\frac{Ri}{Rf}.\left[\left(\frac{1}{A} + 1\right) + \frac{1}{A}.\frac{Rf}{Ri}\right]$$

$$= -v_o.\frac{Ri}{Rf}.\left[1 + \frac{1}{A}\left(1 + \frac{Rf}{Ri}\right)\right]$$

$$\text{Closed-loop gain is } G = \frac{v_o}{v_i} = \frac{1}{-\frac{Rf}{Ri}\left[1 + \frac{1}{A}\left(1 + \frac{Rf}{Ri}\right)\right]}$$

$$= -\frac{Rf}{Ri}\left[\frac{1}{1 + \frac{1}{A}\left(1 + \frac{Rf}{Ri}\right)}\right]$$

$$= -\frac{Rf}{Ri}\left[\frac{1}{1 + \frac{1}{A}\left(\frac{Ri + Rf}{Ri}\right)}\right]$$

Now $B = R_i/(R_i + R_f)$, so the actual closed-loop gain G is :-

$$G = -\frac{Rf}{Ri}\left[\frac{1}{1 + \frac{1}{AB}}\right]$$

For the non-inverting amplifier G is:-

$$G = \left(1 + \frac{Rf}{Ri}\right)\left[\frac{1}{1 + 1/AB}\right]$$

These equations show that the closed-loop gain is reduced if the open-loop gain is reduced.

Loop Gain

The factor AB in these equations is called the loop gain and is an important factor in amplifier calculations.

The error in closed-loop gain due to a finite open-loop gain is :-

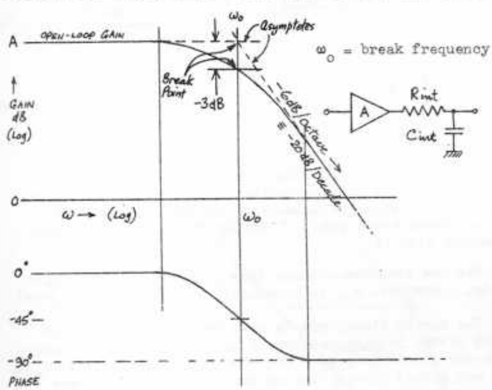
$$\frac{1}{1 + 1/AB} \quad \text{or} \quad \frac{AB}{AB + 1}$$

which is approximately $1 - 1/AB$

If Rf/Ri is greater than unity, B is approximately Ri/Rf and the loop gain A/G. i.e. $AB = A/G$ (A–G in dB)

Bandwidth

For a practical Op. Amp. the open-loop bandwidth is that frequency at which the open-loop gain is 3 dB down on its value at low frequencies and a typical characteristic is shown for an amplifier containing a single time-constant.



The final slope is −6 dB/octave (which is 20 dB/decade) and is known as the asymptote. The 3 dB frequency, which is at the junction of the −6 dB/octave and initial asymptotes, is ω_o and is known as the break frequency or break point. It is determined mainly by stray capacity, Miller effect, etc.

The diagram can be simplified to the asymptotes only and is then called a Bode plot. Such a reducing response is always accompanied by a phase-change - in this case 90° lagging due to capacitor Cs - but transistors themselves do have phase-changes at very high frequencies. With a resistive NFB network the phase, in this example, will not reach ± 180° and the system will not oscillate.

The closed-loop gain depends on open-loop gain as we have already seen and to give an expression for the closed-loop gain of such an amplifier we have only to include a factor for the H.F. loss due to a CR network.

This is :-

$$A_\omega = \frac{A}{1 + j\omega/\omega_o}$$

The closed-loop gain is :-

$$G = -\frac{Rf}{Ri}\left[\frac{1}{1 + \frac{1}{A_\omega B}}\right]$$

which becomes,

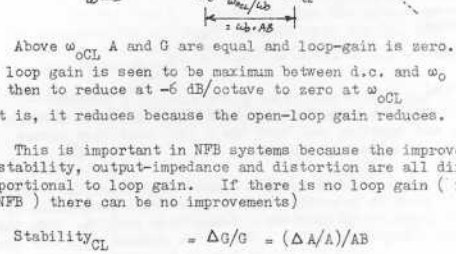
$$G = -\frac{Rf}{Ri}\left[\frac{1}{1 + \frac{1 + j(\omega/\omega_o)}{AB}}\right]$$

The closed-loop 3 dB frequency, ω, is obtained by equating the real and imaginary parts of the denominator :-

$$\text{i.e.} \quad \frac{AB}{AB} + \frac{1}{AB} = \frac{\omega/\omega_o}{AB} \quad \text{or} \quad AB + 1 = \omega/\omega_o$$

$$\text{so} \quad \omega = \omega_o\,(1 + AB)$$

This shows that the bandwidth of the closed-loop gain is greater than the open-loop bandwidth by an amount $\omega_o.AB$, or the original bandwidth multiplied by loop gain.



Above ω_{oCL} A and G are equal and loop-gain is zero. The loop gain is seen to be maximum between d.c. and ω_o and then to reduce at −6 dB/octave to zero at ω_{oCL}. That is, it reduces because the open-loop gain reduces.

This is important in NFB systems because the improvements in stability, output-impedance and distortion are all directly proportional to loop gain. If there is no loop gain (i.e no NFB) there can be no improvements)

$$\text{Stability}_{CL} = \Delta G/G = (\Delta A/A)/AB$$

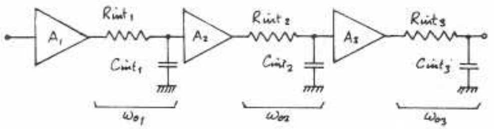
$$\text{Output-impedance}_{CL} = Zo_{OL}/AB$$

$$\text{Distortion}_{CL} = D_{OL}/AB$$

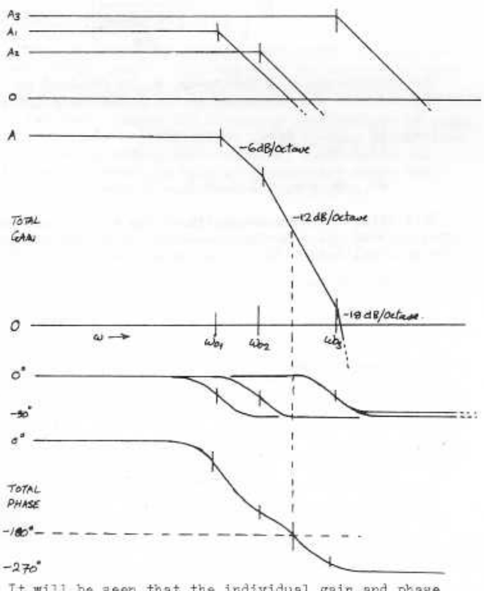
Simple Lag Compensation

Unfortunately, multi-stage amplifiers such as Op. Amps. do not have just one break frequency - they have many; and as each has a 90° phase-shift capacity, the amplifier is virtually certain to have a gain greater than unity where its phase-shift is ± 180° (the conditions for oscillation) even with resistive NFB.

A multi-stage amplifier response is depicted below :-

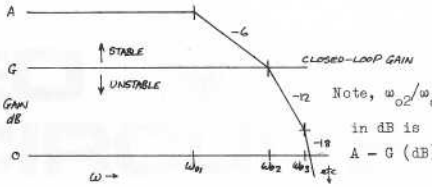


Designers of Op. Amps. endeavour to place the natural breaks due to stray capacities, etc., in ascending order through the amplifier so as to keep the distortion and noise problems to a minimum. Such an amplifier gives rise to the following diagrams.



It will be seen that the individual gain and phase curves add together and that the ± 180° condition occurs for the -12 dB/octave slope.

If a line representing a closed-loop gain, G, is drawn on this plot to intersect the curve of open-loop gain, A, at any place other than the - 6dB/octave portion, then the amplifier may oscillate. This is a general rule for the stability. G must not cross A with a junction slope of more than 6 dB/octave.



Note, ω_{o2}/ω_{o1} in dB is A - G (dB)

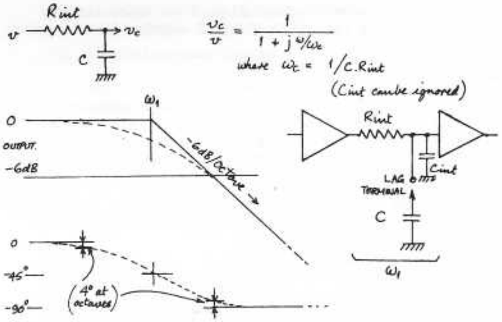
Thus G has a minimum value which is still a considerable gain. This is clearly inconvenient as the most useful Op. Amp. functions are obtained when the gain is around the unity level, 0 dB.

How do we get rid of this excess gain ?

The easiest method is to shunt away the very high frequencies with a single capacitor and this is the principle of frequency compensation.

Frequency compensation enables the closed-loop gain to be reduced at the expense of the bandwidth of the open-loop gain and this is effected by adding a capacitor to a certain point in the amplifier - usually a base circuit - to form a known break frequency or a phase-lagging characteristic.

The characteristics of a CR network are shown below :-



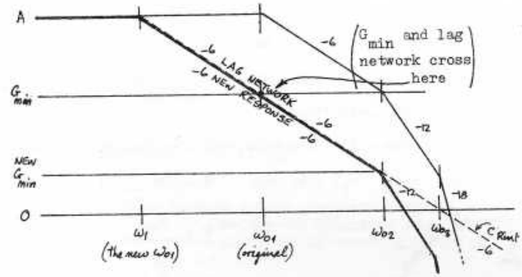
The effect on the Op. Amp. response is shown in the next diagram. To get the optimum effect the -6 dB/octave slope of the capacitor network is arranged to coincide with the first break point of the Op. Amp., ω_{o1}, where G_{min} intersects with it.

The new response extends to ω_{o2} because ω_{o1} no longer exists. However, ω_{o2} is unchanged. (although A is lower)

The curves simply add to give the total gain. The effect of the lagging network is to reduce open-loop bandwidth and closed-loop gain (hopefully to below 0 dB) at ω_{o2} & does not affect closed-loop bandwidth but does increase loop gain below ω_{o1}.

C_{min} can be calculated from the expression :-

$$1/CR_{int} = \omega_c = \frac{\omega_{o1}}{1 + AB}$$

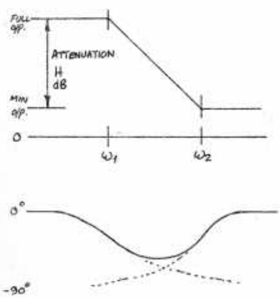


For more feedback and less closed-loop gain C will be larger.

Unfortunately however, the closed-loop gain G is still likely to be greater than unity by quite a considerable amount. Luckily there is a better way to overcome this.

Maximum-Bandwidth Lag Compensation

This uses part of the response of the Op. Amp. to attenuate the loop gain by putting a series resistor with the compensation capacitor.

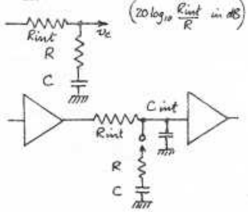


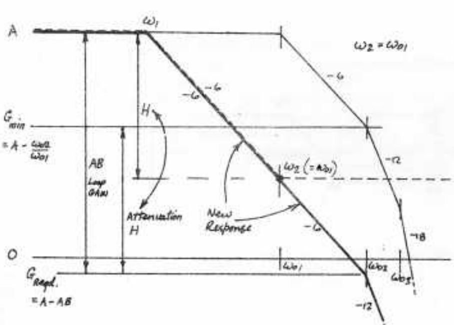
$\omega_1 = 1/CR_{int}$

$\omega_2 = 1/CR$

Attenuation $H = \frac{R + R_{int}}{R}$

or R_{int}/R if atten. is large $\left(20\log_{10}\frac{R_{int}}{R}\text{ in dB}\right)$





The phase-shift of the combination returns to zero above ω_2. The extra attenuation of this method is arranged so that the original second break point, ω_{o2}, is pushed below the required value of closed-loop gain whilst ω_2 is made the same as the original first break frequency ω_{o1}.

The values of C and R are found as follows :-

The required attenuation, $H = \frac{\omega_2}{\omega_1} = \frac{R + R_{int}}{R}$ (dB)

which is approximately R_{int}/R if the attenuation is large.

But, $H = G_{min} - G_{reqd}$ and $G_{reqd} = A - AB$

so $H = G_{min} - (A - AB)$. However, $G_{min} = A - \omega_{o2}/\omega_{o1}$

so $R_{int}/R = A - \omega_{o2}/\omega_{o1} - A + AB = AB - \omega_{o2}/\omega_{o1}$

$$\text{Thus} \quad R = \frac{R_{int}}{AB - \omega_{o2}/\omega_{o1}}$$

C is found from $\omega_2 = 1/CR$, and since $\omega_2 = \omega_{o1}$,

$$C = 1/\omega_{o1}.R$$

General formulæ which allow for phase margins, etc., are :-

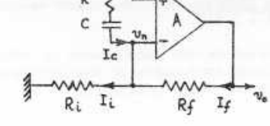
$R = 20(1 + Rf/Ri)\ \Omega$ $\qquad$ $C = (1 + Rf/Ri)/100\ \mu F$

This method of frequency compensation gives the same maximum closed-loop bandwidth as before but enables G_{min} to be made whatever level desired - at the expense of open-loop bandwidth. The original -6 dB/octave slope and the added slope are made into one long slope of -6 dB/octave (to ω_{o2}).

Input Lag Compensation

Simple and Maximum-bandwidth lag compensations both have the effect of slowing down the Op. Amp. in such a way that maximum output cannot be obtained at h.f. no matter how hard it is driven. This is due to the lag capacitors being charged and discharged from relatively high-impedance sources within the Op. Amp. One way to overcome this problem is to put the compensation elsewhere - at the input. This also has the advantage that the compensation no longer depends upon the Op. Amp. and the values of R and C can be made constant over a range of closed-loop gain.

Consider a non-inverting amplifier with R and C added between the input terminals of the Op. Amp. :-



$Rp = \frac{Ri.Rf}{Ri + Rf}$

$B = \frac{Ri}{Ri + Rf} = \frac{Rp}{Rf}$

Some basic relationships are :-

$Ii = Ic + If, \quad Ic = \dfrac{v_i - v_n}{Rp + R + 1/j\omega C}, \quad If = \dfrac{v_o - v_n}{Rf},$

$v_n = Ii.Ri = Ri(Ic + If), \quad v_o = A.v_n,$

Also $v_o = A.Ic(R + 1/j\omega C)$ i.e. A times the voltage between the two input terminals.

$$\text{So} \quad v_n = Ri.Ic + \frac{Ri}{Rf}.(v_o - v_n)$$

$$= Ic.Ri + v_o.\frac{Ri}{Rf} - v_n.\frac{Ri}{Rf}$$

$$v_n + v_n.\frac{Ri}{Rf} = Ic.Ri + v_o.\frac{Ri}{Rf}$$

$$v_n\left(1 + \frac{Ri}{Rf}\right) = Ri\left(Ic + \frac{v_o}{Rf}\right)$$

$$\text{so} \quad v_n = \frac{Ri\left(Ic + \frac{v_o}{Rf}\right)}{1 + \frac{Ri}{Rf}} = \frac{Ri.Rf\left(Ic + \frac{v_o}{Rf}\right)}{Ri + Rf}$$

$$= \frac{Ri(Ic.Rf + v_o)}{Ri + Rf} \quad \text{Hence } v_n = B(Ic.Rf + v_o)$$

but $Ic = \dfrac{v_i - v_n}{Rp + R + 1/j\omega C}$ or $Ic(Rp + R + 1/j\omega C) = v_i - v_n$

$$\text{So} \quad Ic(Rp + R + 1/j\omega C) = v_i - B(Ic.Rf + v_o)$$

$$= v_i - B.Ic.Rf - B.v_o$$

$$Ic\left(Rp + R + 1/j\omega C + B.Rf\right) = v_i - B.v_o$$

Now $v_o = A.Ic(R + 1/j\omega C)$ or $Ic = v_o/A(R + 1/j\omega C)$

and $B.Rf = Rp$, so :-

$$\frac{v_o}{A(R + 1/j\omega C)}.(2Rp + R + 1/j\omega C) = v_i - B.v_o$$

$$v_o\left(\frac{1}{A(R + 1/j\omega C)}.(2Rp + R + 1/j\omega C) + B\right) = v_i$$

$$\text{Gain} = \frac{v_o}{v_i} = \frac{1}{\frac{1}{A(R + 1/j\omega C)}.(2Rp + R\ 1/j\omega C) + B}$$

$$= \frac{A.\left(\frac{R + 1/j\omega C}{2Rp + R + 1/j\omega C}\right)}{1 + A.B\left(\frac{R + 1/j\omega C}{2Rp + R + 1/j\omega C}\right)}$$

Comparing this with $\dfrac{A}{1 + A.B}$ which is the expression for the gain of an amplifier, there is a modifying factor :-

$$\frac{R + 1/j\omega C}{2RP + R + 1/j\omega C}$$

This factor is evaluated in the same manner as before by equating the real and imaginary parts of the denominator, but in this case there are two break points.

$\dfrac{1}{\omega C} = 2Rp + R$, or R, so $1/\omega_1 C = 2Rp + R$

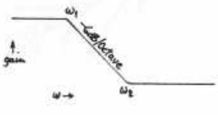
and $1/\omega_2 C = R$

Therefore $C = 1/\omega_1(2Rp + R)$ or $C = 1/\omega_2 R$

for the upper and lower points.

Now $1/\omega_1(2Rp + R) = 1/\omega_2 R$ so $\omega_2 R = \omega_1 2Rp + \omega_1 R$

and $R(\omega_2 - \omega_1) = \omega_1.2Rp$, $\quad R = 2Rp.\dfrac{\omega_1}{\omega_2 - \omega_1}$



$$R = \frac{2Rp}{\frac{\omega_2}{\omega_1} - 1}$$

It will be seen that these two formulae are not dependant on the Op. Amp. As in the case of the maximum-bandwidth lag compensation ω_2 should be put at the original first break frequency ω_{o1}.

ω_2/ω_1 in dB gives the necessary attenuation in dB required to lower the minimum stable closed-loop gain to the desired value.

Hence R is also $2Rp/(H - 1)$

The values for a 702 (see also further on) are given as examples :-

G_{min} is +40 dB, for a gain of 0 dB, H = 40

For a gain of 0 dB, Ri = Rf = 10 KΩ. so R ≏ 10,000/40 = 250 Ω

(the actual value is 220 Ω to give a phase margin)

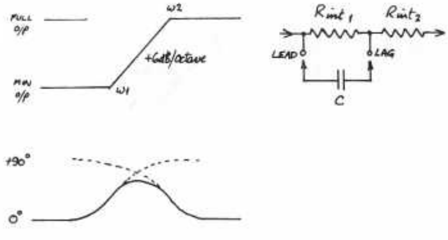
$$C = 1/\omega_2 R = 1/\omega_{o1} R = 1 \Big/ 4 \times 10^6.250 = 1 \text{ nF}$$

This method of compensation has two disadvantages when used with voltage-followers. A value of Rf has to be included, and at h.f. the very high input-impedance becomes low as C becomes low impedance.

Lead Compensation

Most of the higher-performance Op. Amps. have two terminals between which may be connected a capacitor to give phase advance or phase lead.

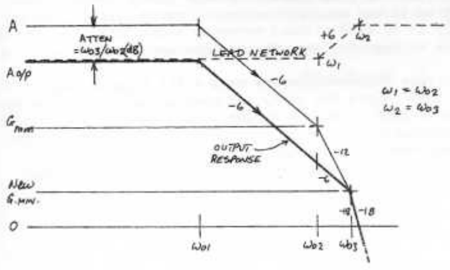
This requires a split series resistor between stages within the Op. Amp. The ratio of R_{int2} to R_{int1} sets the attenuation.



ω_1 is placed at the second break frequency of the Op. Amp. and the required attenuation is ω_{o3}/ω_{o2} (dB). This places ω_2 at ω_{o3}.

Hence $C = 1/\omega_2.R_{int} = 1/\omega_{o3}.R_{int}$ which is small.

The effect is to make a long slope of -6 dB/octave from ω_{o1} to ω_{o3} and to push the -12 dB/octave slope up to ω_{o3}. In fact, there is no -12 dB/octave slope at all.



This extends the bandwidth in the same ratio as the open-loop gain is reduced. The new closed-loop gain can be lower by twice this amount and the loop gain greater.

This method of compensation gives a wider closed-loop bandwidth than that provided by lag compensation, but a compromise between this and a higher loop gain (and less noise) is possible by using smaller amounts of both lead and lag compensation together.

It is not easy to achieve satisfactory performance with lead compensation alone as ω_1 has to be placed with fair accuracy and this is not easy at the higher frequencies. It is all too easy to get part of the -12 dB/octave slope above $G_{min.}$ if C is made too small.

Phase margin

As a guide to a typical Op. Amp. response, the 702 has the following values :-

ω_{o1} 800 KHz, ω_{o2} 4 MHz, ω_{o3} 40 MHz, and -180° occurs at 14 MHz with a gain of +35 dB. Lf. gain is 70 dB so that the loop gain maximum is also 35 dB before oscillations can occur. In actual practise the loop gain cannot be made higher than 20 dB without causing a response peak in the closed-loop gain due to an inadequate phase margin. This means that -6 dB/octave slope is not approached closer than some -3 dB/octave (it is after all a smooth curve and not an asymptote in reality)

Practical compensation (using 702)

In all the previous explanations the phase margin was ignored (i.e. it was zero) but in practise a phase margin of less than 90° will cause a peaked response at the attenuation point. A satisfactory margin value is 45° giving a +3 dB peak. This corresponds to a loop gain of 20 dB with the 702.

With simple lag compensation, 0.1 µF to earth gives a 0 dB bandwidth of 500 KHz and ω is at 250 Hz whilst **A** is 70 dB. This form of compensation is thus ideal for l.f. integrators or other l.f. applications, but little else.

With maximum-bandwidth lag compensation the 0 dB bandwidth is raised to some 4 MHz. ω_1 is now 50 KHz and ω_2 800 KHz. i.e. a four-octave improvement giving 24 dB of attenuation. C is 5 nF and R 40 Ω.

For a 20 dB closed-loop gain C is 1 nF and R 220 Ω giving a bandwidth of some 6 MHz.

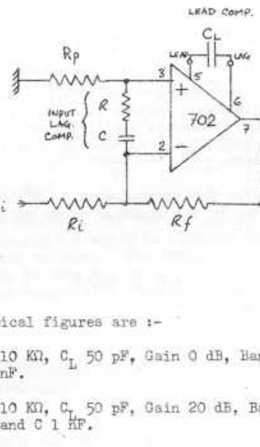
For 40 dB gain C is 100 pF and R 2.2 KΩ and the bandwidth is 7 MHz.

Generally for the 702 the values of C and R are given by :-

$$R = 20\left(1 + \frac{Rf}{Ri}\right) \quad \Omega$$

$$C = 0.01/(1 + Rf/Ri)\ \mu F$$

With the 702 using lead compensation, the closed-loop bandwidth is some four-times greater than using lag compensation only, but some lag compensation is required below 40 dB closed-loop gain.



$Rp = \frac{Ri.Rf}{Ri + Rf}$

Typical figures are :-

Rf 10 KΩ, C_L 50 pF, Gain 0 dB, Bandwidth 23 MHz, R 200 Ω, and C 1 nF.

Rf 10 KΩ, C_L 50 pF, Gain 20 dB, Bandwidth 26 MHz, R 200 Ω and C 1 nF.

Rf 10 KΩ, C_L 50 pF, Gain 40 dB, Bandwidth 28 MHz, R and C not required.

A low-priced Op. Amp. with no built-in compensation is the 702. This has two terminals for compensation purposes which are labelled LEAD and LAG.

With some other types of Op. Amp. (e.g. 709) the lag compensation capacitor is not earthed but returned to another point in the amplifier.

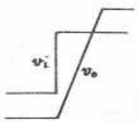
Load Capacity

The capacity of the load on the output of the Op. Amp. has an effect on the phase of the NFB at h.f. and the lag due to this is effectively added to the phase curves for the Op. Amp.

It can be minimised by increasing the phase margin by increasing C and decreasing R (both by a factor of two or more) or by putting a small capacitor across Rf to give some phase lead to effect cancellation.

Slew-rate

Not the mortality-rate of Op. Amps., but an expression for the maximum rate-of-change at which the output of the Op. Amp. can change in response to a step-function input. - i.e. the gradient of the slope.



The signal rate-of-change is directly proportional to the amplitude but the Op. Amp. with lagging capacitors has to charge and discharge these at the signal rate and there is a limit to each particular point in terms of charging current. If this rate is exceeded then the Op. Amp. cannot catch up with the signal until the signal slows down again. This effect is expressed in Volts/µs.

At high frequencies then, there has to be a compromise between amplitude and frequency - you can't have both.

The slew-rate and the bandwidth are often quoted separately from the full-output maximum frequency which can be very mis-leading. Likewise the slew-rate is often quoted at a high gain figure instead of at unity gain where it is much less.

Maximum-frequency for full output, with no distortion, and slew-rate are related by the equation shown below (the gain being unity) :-

$$\text{Slew-rate } S = \frac{d\,v_o}{dt}_{max} = \omega.V = 2\pi f.V$$

Noise and compensation

The compensation network affects the noise performance as lagging compensation attenuates the h.f. components of both noise and signal.

Input lag compensation actually worsens the signal-to-noise ratio since the network does not attenuate Op. Amp. noise - only signal h.f. A factor of 20 dB is typical. At unity gain this may be even more.

Noise and Slew-rate performances therefore are both dependant on compensation in opposite ways and a compromise may have to be reached in some applications by using both types of compensation.

A typical performance of the 702 is given:-

Lagging comp., Bandwidth 10 KHz, Slew-rate 0.35 V/µs,
Noise 15 µV.

Input comp., Bandwidth 800 KHz, Slew-rate 30 V/µs,
Noise 170 µV.

(both at unity gain)

The 741 has a slew-rate of 0.5 V/µs which gives a unity gain bandwidth of 22 KHz. This is not much use for T V work as it is generally assumed that some ten harmonics are necessary to reproduce a waveform faithfully

This really confines the use of the 741 to d.c. and l.f.. It is difficult to get one to handle a line-rate sawtooth waveform with any satisfaction.

The 741 is a very useful Op. Amp. though despite this drawback as it has no compensation to worry about. Quite literally it is used without worrying about other components; as in the voltage-follower where there are indeed none at all.

All this compensation may seem a bit frightening but it is made very simple for us by the manufacturer. All we have to do is to look up in his tables our value of G and read off the corresponding values of C and R for the different types of compensation. Of course, there is usually so much information and so many symbols that this is not all that easy. But that is another problem...........

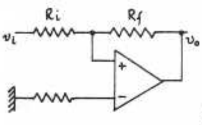
Schmitt Trigger

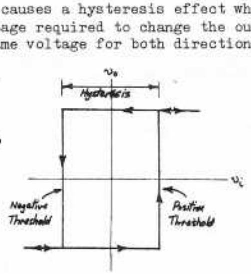
So far in this series on Integrated Circuits the Op. Amp. has been considered only with Negative Feedback, NFB. What is the effect of Positive feedback ?

This is of course feedback from the output to the non-inverting input. With no feedback at all the very high gain of the Op. Amp. usually ensures that the output will be in a saturated state due to inherent imbalances in the input circuitry. There is only a small region of input voltage near Ov where the amplifier can in fact amplify - this is $(+V_{sat} - -V_{sat})/A$. It should be remembered that Op. Amp. supplies are normally two-rail or positive and negative.

Positive feedback reduces this voltage range to virtually nothing so that the output is always in one saturated state or the other and the input voltage range can be considered to be a single voltage. Positive feedback does this by assisting the input voltage change - causing regeneration. A circuit of this type is known as a Schmitt-trigger or level-detector.

The addition of PFB causes a hysteresis effect which means that the input voltage required to change the output state is no longer the same voltage for both directions of change





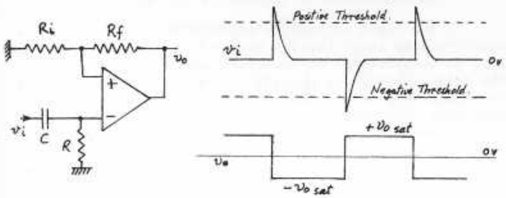
The voltage at which the change occurs is called the threshold and has the value $\pm Vo_{sat}\, Ri/(Ri + Rf)$.

The circuit is useful for sharpening up waveforms such as slow pulses and sine waves.

Simple Bistable or Latch

The schmitt-trigger circuit forms the basis of a simple bistable in which a differentiating input circuit is used. This passes only the edges of the input pulses and these make the Op. Amp. change state if their voltages exceed the threshold levels. The mean level is 0v which is within both threshold levels and so to change states alternately the trigger pulses must also be alternate in polarity.

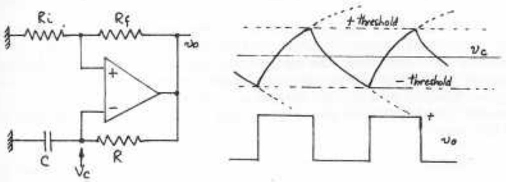
The circuit is useful as a latch or memory.



OSCILLATORY CIRCUITS

Relaxation oscillator or Multivibrator

A schmitt-trigger type of circuit is partway towards being an oscillator - all that is required is a time-constant to delay the switching action in a predictable manner.



The operation is that the Op. Amp. saturates at some voltage $\pm Vo_{sat}$. The voltage at the non-inverting input is then determined by the feedback ratio and is $\pm Vo_{sat} \cdot \frac{Ri}{Ri + Rf}$

If the Op. Amp. is in positive saturation then the fed back voltage is also positive and so C charges positively. via R until Vc reaches $+Vo_{sat} \cdot Ri/(Ri + Rf)$.

The amplifier output then changes state because the inverting input starts to go more positive than the non-inverting input. That is , the output goes negative. C then begins to charge towards $-Vo_{sat} \cdot Ri/(Ri + Rf)$ and so on.

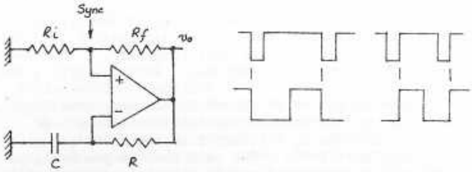
The period of oscillation is:-

$$2CR \log_e \frac{1 + Ri/(Ri + Rf)}{1 - Ri/(Ri + Rf)} = 2CR \log_e (1 + 2Ri/Rf)$$

If Ri = Rf then t = 2.2 CR seconds.

The oscillator can be synchronised by inserting narrow pulses into the non-inverting input.

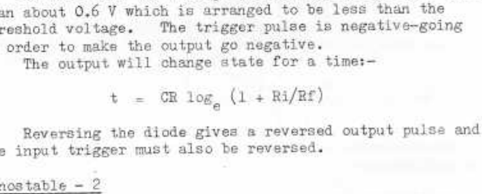
These pulses may be either positive or negative-going and will synchronise positive and negative edges respectively. It is generally preferable to use only one polarity.



The frequency stability with supply voltage is good - unlike the common transistor multivibrator - because if the supply is, say, halved the the charging current is also halved but so too is the threshold voltage. The switching voltage is therefore reached in about the same time as previously.

Monostable - 1

The multivibrator may be turned into a monostable, or one-shot, merely by adding a single diode across the capacitor. This prevents the circuit from being self-acting in one direction.



In this case the voltage across C cannot go more positive than about 0.6 V which is arranged to be less than the threshold voltage. The trigger pulse is negative-going in order to make the output go negative.

The output will change state for a time:-

$$t = CR \log_e (1 + Ri/Rf)$$

Reversing the diode gives a reversed output pulse and the input trigger must also be reversed.

Monostable - 2

Another type of monostable can be made in which the period is determined by a reference voltage.

The timing capacitor is placed in the PFB path and the output is normally at $+Vo_{sat}$. The inverting input will be at -Vref (as is the other input).

The input trigger pulse must exceed Vref in order to unsaturate the Op. Amp. i.e. it must be positive-going with an amplitude greater than Vref. The output voltage change from + Vsat to - Vsat is passed via C to the non-inverting input and C charges via R until the voltage at the non-inverting input equals Vref. (Vref must be less than -Vo sat)

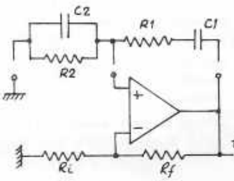
The period of the output is $CR \log_e(2Vo_{sat}/Vref)$

and the output period may be longer or shorter than the input trigger pulse.

Sine-wave Oscillator

Sine waves can be obtained from an Op. Amp by using a Wien Bridge network. This has an attenuation of 3 times and a zero phase-shift at the null frequency. The frequency at resonance is given by $1/2\pi RC$ Hz. Atten. = $1/(1 + 2R1/R2)$

The network has to be placed in the non-inverting feedback loop and to maintain oscillation the gain of the Op. Amp. must be 3, i.e. Rf/Ri in the inverting feedback loop must be 2. The gain must be precisely 3 otherwise the oscillator either will not start or will saturate. One way of stabilising the gain at 3 is to use a thermistor for Rf (the R53 by STC is suitable)



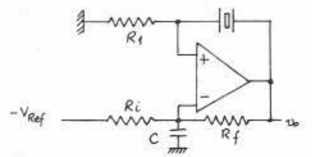
← Wien Network

Rf = 2Ri

For 1 KHz and 3V output pp, R=10KΩ and C= 16 nF, Ri=470Ω and Rf = R53(thermistor). Both amplitude and frequency are stable to within 0.1 %.

Crystal Oscillator

The crystal is used to provide positive feedback and resonates in its series mode - approximately 0.1 % low in frequency. The series capacitor will enable the frequency to be raised to the correct frequency.



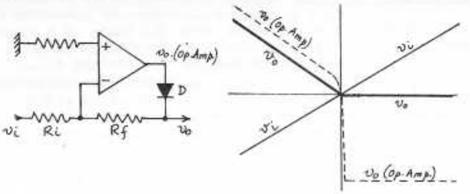
The following conditions should be satisfied for good performance :-

R1 = Ri/Rf/(Ri + Rf), Rf ≙ 39 KΩ, Rf/Ri ≙ 1.35 and Xc ≙ Rf/500. CV should be about 5 pF for a 5 MHz Xtal.

MORE USEFUL CIRCUITS - Rectifiers

The high gain of an Op. Amp. can be put to use in a half-wave rectifier to reduce the forward drop due to a semi-conductor. For a silicon diode this is about 0.6 V and for a Germanium diode about 0.3 V. If low voltage signals are to be rectified, as in most detectors for example, then the knee of the diode curve causes considerable distortion, because the output is not linear in this region. The ideal diode has no forward voltage drop of course.

Placing the diode in the feedback loop of an Op. Amp. effectively divides the forward drop by the open-loop gain. and for most purposes it is then negligible. Very low amplitude signals can then be detected with little distortion.



If the input voltage is negative with respect to the non-inverting input (earth) then the output of the Op. Amp. must be positive and so the diode must be conducting. The output voltage is exactly the inverse of the input as the diode drop is overcome by the Op. Amp. output being that much more positive than the output point. The forward resistance of the diode plus Rf makes the gain of the Op. Amp. greater than unity but the output is attenuated by the same amount due to the inverting input being a low impedance. Hence the gain is Rf/Ri as usual. It is seen, therefore, that the diode drop is eliminated.

If the input is positive the the diode is non-conducting and the output point potential is that of the inverting input which must be that of the non-inverting input, or 0V.

The output point thus has two values of impedance - low or Rf. A voltage-follower should therefore ideally follow a half-wave rectifier.

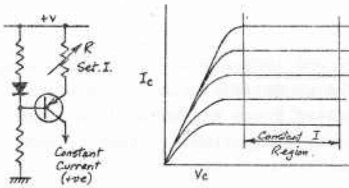
A number of half-wave rectifiers can be joined together in the usual manner to form full-wave rectifiers, etc., merely by commoning the output points. Such a circuit gives an output which is the lowest of the inputs. In other words, a non-additive mix. This system can be useful in a pattern generator, and is in fact an analogue NAND gate.

Constant-current source

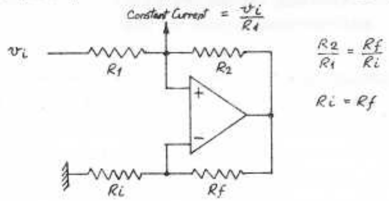
A constant-current source is one whose impedance is high so that the voltage varies whilst the current does not when a variable load is present. One of the most common uses for such a source is to charge capacitors so as to form a linear sawtooth waveform. Another is to form the 'tail' of a long-tailled pair. This enables the gain to be kept high.

The normal method of making a constant-current source is to use a transistor as shown. This has some disadvantages amongst which are the use of a diode to compensate for the forward drop of the transistor which varies with temperature, and the fact that the current has to be set with a variable resistor R.

The Op. Amp. gives us a convenient voltage-variable source which is temperature-independant and as an added advantage can be given an adjustable impedance.



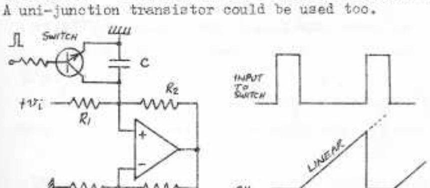
The Op. Amp. can be consisdered to have a gain of +2 at the non-inverting input since the NFB resistor is the same value as the input resistor. On the other hand there is positive feedback to the same input which is attenuated by 2 because of R2 and R1. Any change at the non-inverting input is thus accompanied by an identical change at the same point due to the Op. Amp. This can be considered as a resistor with the same voltage at both ends - it has infinite resistance. The non-inverting input thus exhibits infinite resistance which is the condition for a constant-current source. Another name for this principle of 100 % feedback is bootstrapping.



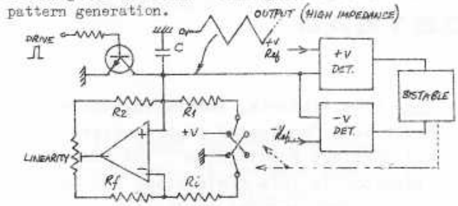
A load connected to the non-inverting input will receive a current of +V/R1 and although the voltage across the load may change, its current will not. The maximum input voltage is Vsat/2 because the gain is two. The input voltage could be applied via the inverting input instead and would then give a negative current.

This circuit provides a convenient way of generating sawtooth waveforms as the next diagram shows. The transistor is used to discharge the capacitor to start the sawtooth. The voltage across the capacitor is directly proportional to the current in R1, which is proportional to +V, and so a positive ramp results as C charges up.

The sawtooth level may be detected at a certain level by means of a schmitt-trigger or similar level-dependant circuit and the output used to turn on the discharging transistor. This results in a free-running generator. A uni-junction transistor could be used too.



If a bistable is driven from the voltage-detector the output can be used to change inputs to the constant-current source so that the ramp, instead of being taken down to OV by the switch, descends at the same rate as it ascended. Another voltage-detector is then usеd to reverse the bistable again, and so on. This generates a triangular waveform. This generator can be used for pattern generation.



It is evident that this form of constant-current source is very useful and it has two more advantages yet. The first is that either positive or negative currents may be obtained by positive or negative voltage (low impedance) to R1 or by negative or positive voltage (low impedance) to Ri. The second is that by varying the ratios of the feedback arms the linearity of a sawtooth waveform can be varied in both directions. Furthermore, the linearity control is to a large extent independant of the amplitude, i.e. it does not cause an amplitude change. This system of sawtooth generation is therefore ideal for a scan generator.

Errata

Two diagrams were interchanged in CQ-TV 80 in part 10 of this series. These were the last diagram on page 10 and the first diagram on page 11.

Also in CQ-TV 80 p10, the differentiator gain-limiting resistor Ri should be in series with Ci - not parallel.

Next issue

The next part of this series will examine some more uses of Op. AMps., also better Op. Amps. than the 741 and will include some other types of Linear ICs which have been held over from this time.

It is also hoped to describe the 7-segment decoders and indicators in the next issue, as well as some more digital circuitry.

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Acknowledgement

The author wishes to thank the Directors of EMI Sound and Vision Equipment Division for permission to publish this article.

POSTBAG

Chris Long from Victoria, Australia, describes himself as a "dedicated experimenter in the field of video electronics". He has been interested in this field since the age of 14 and his experimental studio now includes, among other interesting equipment, a range of mechanical scanning equipment for low definition tv. These are all based on a 32 line standard, and a recent 30 line Baird system recording made by Chris on ¼ inch tape is now being used by the I.B.A. in their "Television Gallery". Although interested in high definition tv (he has built a 1,500 line flying spot scanner) the main interest is low definition television (L.D.T.V.) and a recent extension of this work was a plan to transmit moving 30 line pictures from Australia to Britain. Although supported by the Wireless Institute of Australia, the plan was unfortunately frustrated by what Chris' English colleagues describe as "condescending opposition by the T.S.G.B." in Britain. It is now found much less restrictive to carry out "transmissions" by means of ¼ inch tape sent through the post. An attempt is now being made to whip up some more interest in L.D.T.V. and to get some more people to become involved in tape exchanges.

Franklyn Brooker 9Y4VU c/o Instrument Dept, Texaco, Pointe-a-Pierre, Trinidad asks B.A.T.C. members to note that he is available on any band for SSTV contacts.

G. James McKee Jr. WB6ROP from the Southern California ATV Club has just joined B.A.T.C. and has sent us a recent copy of their monthly Club Bulletin - thanks Jim! The Club has regular meetings - the December one included a tour of a nearby TACAN, VOR and LRCO station with a lecture, followed by an auction of gear. Sounds a well organised group which could be taken as a model for many others. Keep up the good work!

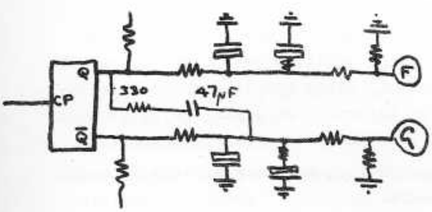
Gareth Evans G8DXY from Bristol has written to point out a ghastly editorial error in C Q - T V 80; the photograph in the Convention report allegedly of Nigel Walker and his colour equipment is actually one of Gareth fiddling with Nigel's colour equipment! In Gareth's own words "I think a small apology is due to Nigel". Too right! The Editor humbly apologises to both of you.

Ladislav Vig from Switzerland has written telling us more of his experiments with PIN Photodiodes for Flying Spot Scanners. He hopes for success along these lines very soon, but has heard that phototransistors, previously thought to be too slow, may now be feasible due to new circuitry developed by Siemens. Ladislav has also used the GW6JGA/T scan circuit published in C Q - T V 78, with some modifications; he has added a current amplifier for the scanning coils, and also reversed the diode D1, resulting in negative going blanking at the output, to suit his own experiment. For a 625 line scan unit he has developed another unit using a faster LM201 op amp.

B.W. Smith ZUCJ from Huntley, New Zealand, is a new member of B.A.T.C., and tells us that he uses the C Q - T V S.P.G. to drive his one inch vidicon camera. The S.P.G. worked first go he says (well, almost, he'd only forgotten one link which had to be resoldered!) and apparently several other people in the area are also building it, one of them being ZL1TFX. As Arthur Critchley points out in his article, this S.P.G. really is being built in large numbers. Let's hope it gives you long service, Brian!

David Wilkinson of New Eltham, London wrote recently with a few topics of interest to readers. First, the Multiburst Generator circuit printed in C Q - T V 80 had due to an oversight, the horizontal cross-straps on the three right hand monostables omitted. They should be strapped as per the left hand one (but not have an output taken from the strap as does the left hand one). Secondly, readers may have seen recently adverts for an I.C., the XR205 Monolithic Waveform Generator; here is some information about it. It contains a voltage controlled multivibrator, a balanced modulator, a buffer amplifier, and more interesting, provision for a sinewave output. This latter facility is presumably achieved by applying the triangular wave output to a suitable diode network. Out of curiosity, an attempt was made to use this as a video sweeper. It was found that the maximum output frequency was about 3.5MHz but only a 3:1 sweep range could be achieved. (At lower frequencies 10:1 can be achieved). Also the output level falls somewhat above 2MHz. In fairness, this performance is in accordance with the data sheet.

On the subject of the C Q - T V S.P.G., mention was made in C Q - T V 80, page 13, of the difficulty with phase modulation when mains locking. This can be greatly improved by adding an anti-phase a.c. component from the other side of the bistable (see Fig. 18, C Q - T V 77, page 11).



The 47 F should be a reversible type. The 330 ohm resistor is best adjusted for minimum modulation. The idea came from a fairly vintage Sylvania S.P.G.

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THE CROFTON CAMERA KIT

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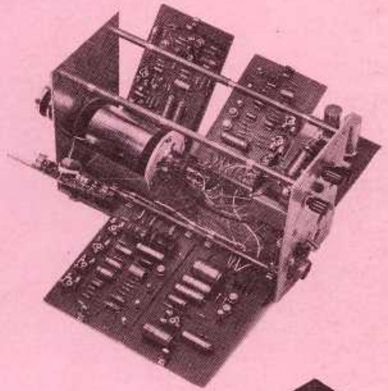

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