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Stereo coder

1—Choice of method/oscillator stability

by Trevor Brook

A practical design for a high quality coder suitable as a test instrument is described. Apart from the audio filtering, inductors have been avoided and a compact board layout produced. A v.h.f. unit for servicing checks on receiver performance could also be used by demonstration showrooms to feed programmes of their own choice to stereo tuners.

Part 1 examines the stereo multiplex system and establishes tolerance limits for signal components. Channel separation is considered as this would assume increased importance if a matrix system of surround sound broadcasting were adopted. Part 2 gives construction and alignment details for the coder and Part 3 gives modifications to the Portus and Haywood decoder to provide a low distortion reference decoder.

Work on this coder started originally out of curiosity as to whether an inductorless design would be possible. Early experiments were promising and the design has been pursued to give performance of broadcast quality.

The specifications of stereo coders now in use at both national and independent local radio transmitters are given in Table 1 and most existing coders have similar figures. Particular objects of this design are to improve crosstalk at the higher audio frequencies and achieve mid-frequency distortion better than 0.05%.

Stereo signal specification

The modulating signal in the Zenith-GE pilot tone system is defined as

$$0.9\left(\frac{A+B}{2}+\frac{A-B}{2}\sin 2\pi f_c t+0.1\sin \pi f_c t\right),$$

where A is pre-emphasized left channel, B is the pre-emphasized right channel, and f_c is 38kHz. $\frac{1}{2}(A+B)$ is called the sum or M signal, and $\frac{1}{2}(A-B)\sin 2\pi f_c t$ is called the stereo difference or S signal, and is a double sideband suppressed-carrier signal. $0.1\sin \pi f_c t$ is the pilot signal at 19kHz.

Substituting the maximum values A = +1 and B = +1 or −1 gives the maximum amplitude of 90% for the M and S signals respectively. Monophonic receivers continue to produce only the M signal as audible output thus giving the system its compatibility.

Decoding

To retrieve the stereo information involves a decoder which can take the form in Fig. 1. The reduction in channel separation if a decoder adjusted to decode a perfect multiplex signal is presented with signals having the five following departures from ideal is shown in Fig 2:

– amplitude error between the M and S signals

–phase error between the M and S signals

–phase error in the pilot relative to the 38kHz suppressed carrier. The requirement for pilot phase accuracy is substantially less than for M/S phase accuracy

–amplitude error of one sideband only of the S signal, typical of the h.f. loss

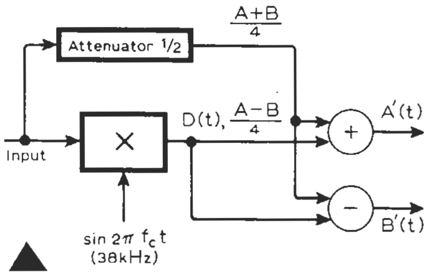


Fig. 1. How the stereo multiplex signal can be decoded. After de-emphasis the A′ and B′ outputs become left and right channels.

Fig. 2. Inherent crosstalk of the multiplex signal plotted against
1) Amplitude error between the M and S signals
2) Phase error between the M and S signals
3) Error in pilot phase
4) Amplitude imbalance between the sidebands of the S signal
5) Phase shift in a sideband of the S signal

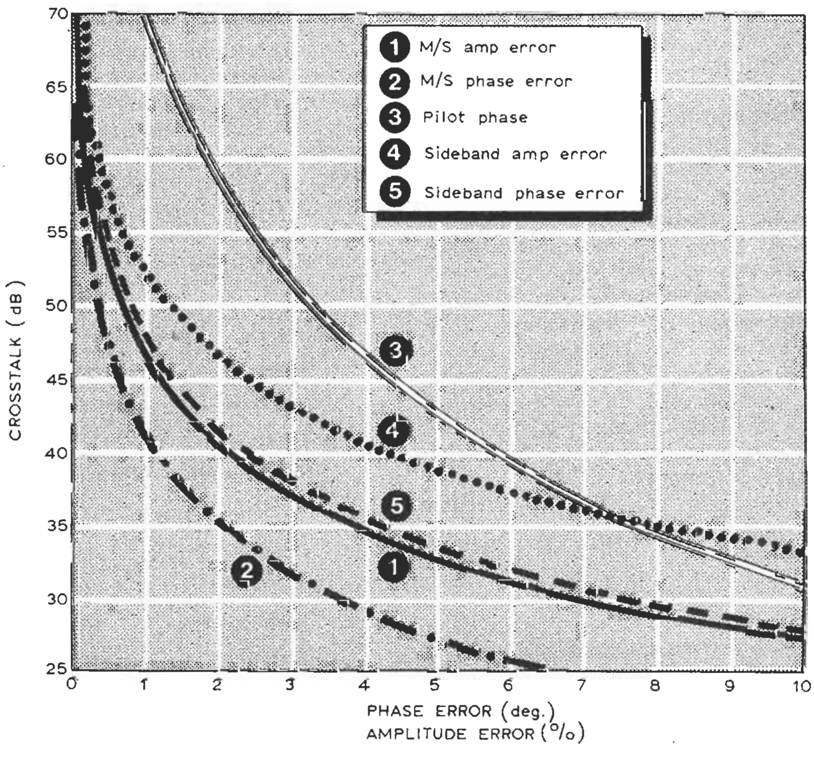


Table 1. Some parameters of the broadcast coders in use today.

	BBC	IBA
Amplitude response ±0.5dB	60Hz — 10kHz	50Hz — 15kHz
± 1dB	40Hz — 15kHz	—
Channel separation	100Hz — 10kHz ⩾40dB	400Hz – 5kHz⩾48dB
	10kHz – 15kHz⩾36dB	30Hz – 15kHz⩾42dB
Harmonic distortion (1kHz)	0.3% at 2dB above peak level	0.5% at peak level
38kHz leakage	<40dB	⩽40dB

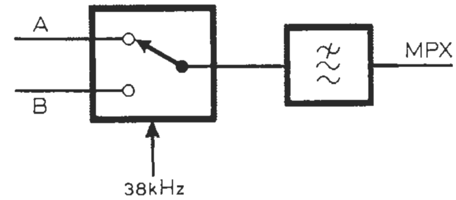


Fig. 4. Basic arrangement for the switch encoding method. Switch would typically be a ring of diodes or f.e.t. switch driven by the 38kHz square wave.

encountered in receivers
–phase error of one sideband only of the S signal.

For reasonably high channel separation, say better than 45dB, the above effects may be considered as algebraically additive. It is evident that extremely stringent amplitude and phase performance requirements are set for a coder intended to give high channel separation.

Another problem in the multiplex system is distortion. Apart from the usual harmonic and intermodulation distortions, spurious beat tones can be produced in the decoded outputs. This is the result of intermodulation between the various components of the stereo signal and, though predominantly a receiver problem, could also be caused in the signal generation method or in a coder's output amplifier. Beat tone distortion is worst at the higher audio frequencies and subjectively produces an unpleasant 'splashing' sound on sibiliants. Most stereo receivers will give clearly audible low- or mid-frequency beats on the 10 and 14kHz bursts during the BBC stereo test zone transmissions even though these tones are not at full level. On mono reception of a stereo signal the effect is not noticeable except on very poor receivers. Fig 3 shows the beat tone possibilities in both mono and stereo reception.

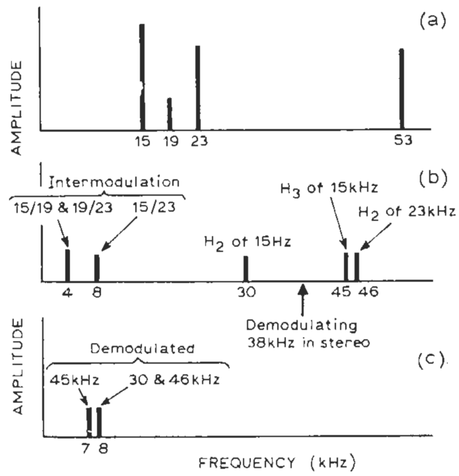


Fig. 3. Principal possibilities for production of beat tone distortion in any receiver (b) whenever a composite stereo signal (a) undergoes intermodulation distortion. Additional tones are produced in a stereo receiver (c).

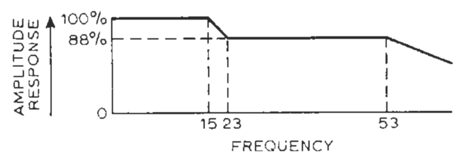


Fig. 5. Desired characteristic for the filter in the simple switch coder, Fig. 5. A linear phase response is required up to 53kHz.

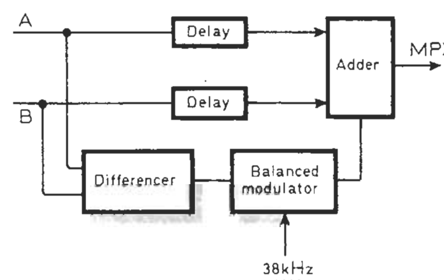


Fig. 6. Block diagram for the matrix coding method.

Generating multiplex signals

There are two principal ways of producing the coded stereo signal. The first and almost universal are switching methods, while the second is the matrix method where the individual signal components are generated separately and then added together.

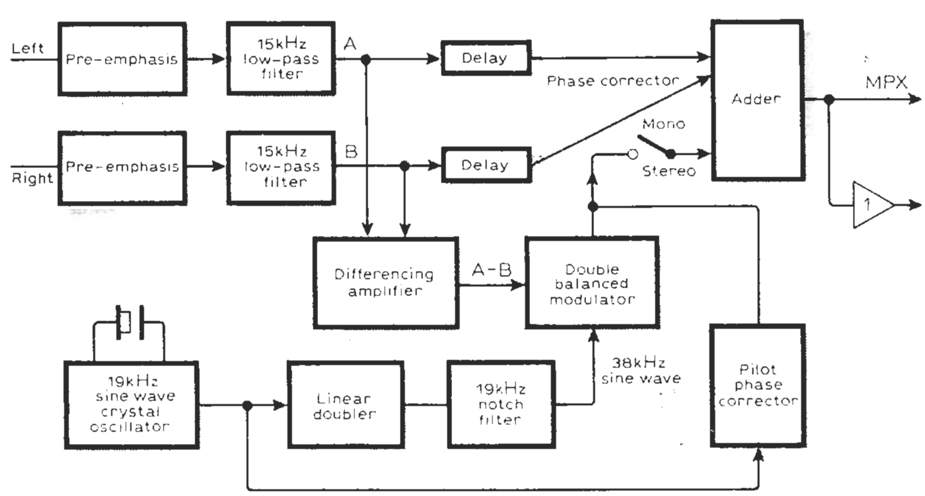
Conceptually the simplest and also a common way of switch encoding is to switch between the A and B signals with a diode ring or similar device driven by 38kHz, Fig 4. For a square wave switching signal the following output is produced

$$\frac{A+B}{2}+\frac{A-B}{2}\cdot\frac{4}{\pi}\left|\sin 2\pi f_c t + \frac{1}{3}\sin 6\pi f_c t+\frac{1}{5}\sin 10\pi f_c t + \ldots\right|$$

The snag is that sidebands around odd harmonics of the switching frequency are present in the output and, more difficult, the required difference signal has too high an amplitude. To remove the components above 53kHz as well as reduce the S amplitude by $\pi/4$ requires a filter with the amplitude characteristic shown in Fig 5, and a linear phase response at frequencies below 53kHz! Broadcast quality coders typically use ten or so inductors in such a filter. Other switching arrangements can avoid the necessity for S amplitude correction and in other respects switching coder performance is largely determined by switching time and 1:1 accuracy of the 38kHz square wave. Whatever low-pass filter is required practical realisations need careful alignment involving several interdependent adjustments and deficiences in the filter make such coders susceptible to all the forms of signal degradation listed earlier. A complete switching coder design has been published by Mack[1] and the virtues of diode cross modulator circuits for applications including coding stereo extolled by the same author[2].

The matrix form of coding is shown in Fig 6. A point to note is that the 38kHz fed to the multiplier in this case is a sine wave. Another alternative would be to use a conventional switching multiplier

Fig. 7. Complete block diagram of the coder described.



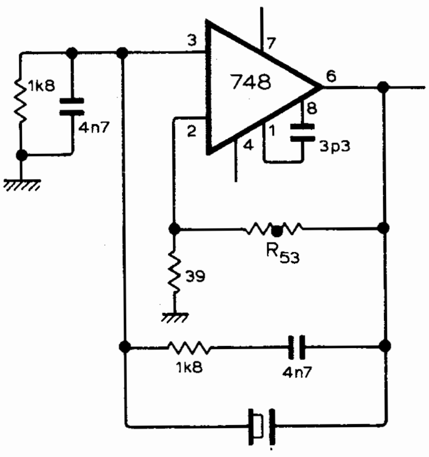


Fig. 8. Thermistor-controlled oscillator with the R53 bead running at 205°C, and a resistance of 82 ohms.

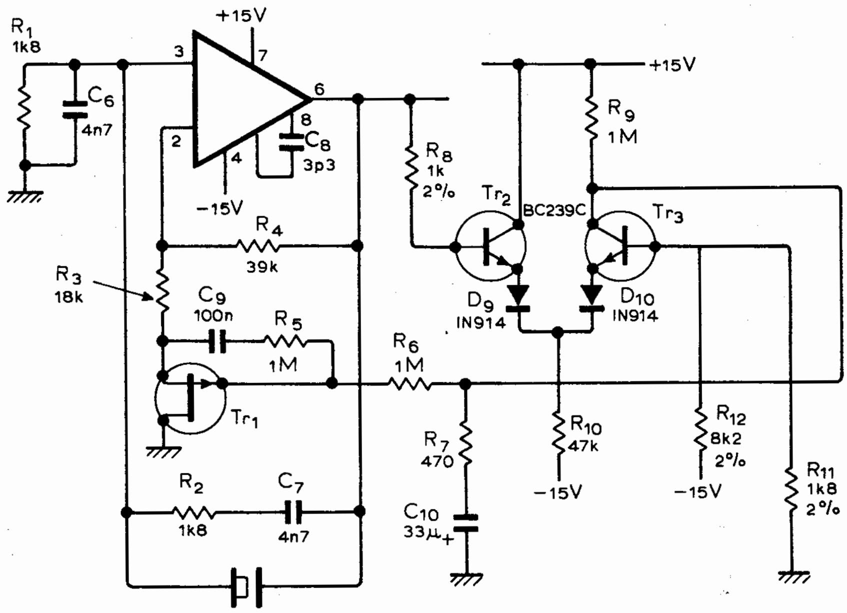


Fig. 9. Oscillator with improved amplitude/temperature drift. Tr_2 and Tr_3 form a long-tailed pair comparator and R_5, R_6 equalize the signal voltages across the f.e.t. to linearize it and reduce distortion of the sine wave output.

fed by a square wave, filter out the difference signal from the odd harmonic components, and then feed it to the adder. With the matrix arrangement in the diagram the last three degradations can be made negligible so crosstalk performance mainly rests on achieving good gain stability and phase matching between the sum and difference signal paths.

However, some new problems arise with this form of coder. The linearity of the multiplier to audio frequencies will have an effect on beat tone distortion performance and in a practical design there is the danger of impurities on the 38kHz producing further beat tone outputs. Common to all forms of coder is the need for a low distortion 19kHz pilot of correct phase and good stability, and audio pre-emphasis and filtering to limit the bandwidth of A and B signals to 15kHz.

Choice of matrix method

With an instinctive loathing of inductors and poor prospects at the time of realising a sensible active filter meeting the stringent phase and amplitude requirements while introducing negligible noise and distortion, the matrix approach looked more promising. Using the matrix principle only the first two signal degradations should be apparent and to meet a target channel separation of 55dB implies an M/S amplitude error of less than 0.18% and an M/S phase error better than 0.1°.

The block diagram of the coder is shown in Fig 7. Both 19 and 38kHz sine waves are required in this coder and starting with a 19kHz sine wave which is then doubled using a linear multiplier to square its input (a sine-wave squared equals a single wave of double frequency plus a d.c. term) and produce 38kHz of correct phase involves less filtering than starting with a square wave at a higher frequency (38, 76 or 152kHz) and dividing down.

If the coder is to be fed from any practical signal source other than a distortionless audio signal generator then the 15kHz filters are essential to prevent gross beat tone effects due to ultrasonic components on the audio inputs. It is desirable that rejection of frequencies of 19kHz and above be at least 45dB so this means that when pre-emphasis is in use, giving around 20dB boost at 19kHz, a filter attenuation of 65dB is required by 19kHz. If a passband ripple of only 1dB is allowed this implies the enormous attenuation rate of 200dB/octave between 15 and 19kHz.

Fortunately two cascaded Toko filter blocks can exceed the requirements in a very small space and at reasonable cost. The drawback of a filter with such violent attenuation so close to the passband is that there is little hope of achieving a linear phase characteristic and this is a deficiency common to all stereo coders. Part 2 includes a spectrum analyser photograph of the present filter response and a graph of measured phase shift.

The audio difference is derived and fed via the balanced modulator to the output adder while the sum signal is produced by feeding equal amounts of A and B directly to the output adder. The longer route of the difference signal means that it is slightly delayed at the adder compared to the A and B components and at the higher audio frequencies this would amount to a significant phase shift between the M and S components, hence the phase correctors inserted in the A and B lines to the adder.

Because the linear doubler used to produce 38kHz is not a perfect device some leakthrough of 19kHz may occur, particularly at extremes of the temperature range, so a 19kHz rejector is placed before the multiplier's carrier input. The pilot must also arrive at the output adder at the correct amplitude and phase and a small phase shift is required to equal the time delay through the doubler, amplifier, notch filter and balanced modulator. To provide balanced outputs a straightforward unity gain invertor is fed from the adder output.

Power supplies are entirely conventional, producing plus and minus 15 volts at around 100mA. Power take off points are provided for running the clipping amplifier and v.h.f. oscillator described later.

19kHz oscillator

The accepted frequency tolerance for the pilot tone is ±2Hz so crystal control, if not essential, is certainly desirable. As a sine wave is required anyway it seems sensible to start with a sine wave crystal oscillator. This is something which often gives circuit designers a problem but a reliable inductorless circuit is easily formed at low frequencies by building a Wien bridge oscillator around the correct frequency and then putting the crystal across the series element of the bridge. Easy starting with reliable crystal lock is the result.

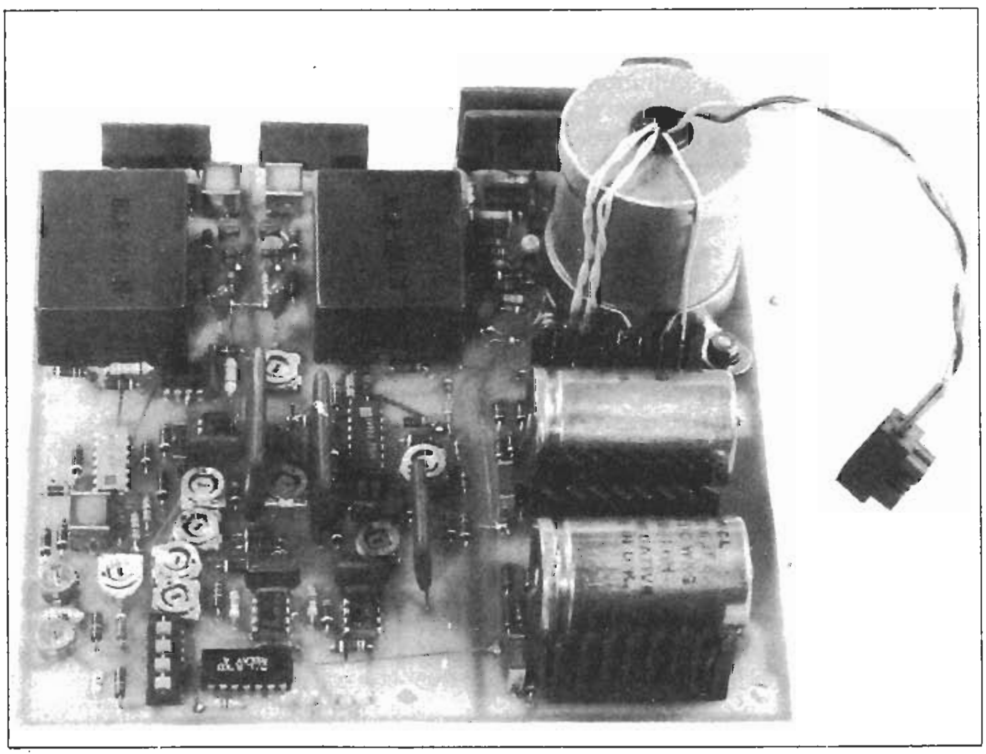
The standard thermistor amplitude control method, Fig 8, proved adequate for an early prototype, but even running the bead as hot as permissible, 200 deg C, still means that its operating point is determined roughly 9 parts in 10 by the oscillation voltage and 1 part in 10 by the ambient temperature. With this circuit I measured an amplitude drift of −0.02dB/deg C over the range +10 to +40 deg C and distortion was below 0.05%. Evidently some form of amplified control was needed to improve this drift figure tenfold and allow maintenance of good channel separation over a wide temperature range.

The circuit evolved is in Fig. 9 where an f.e.t. replaces the thermistor as the gain control element but with linearizing components to maintain the distortion performance. Linearizing is achieved by equalizing the gate/drain and gate/source signal voltages and is done by R_5 and R_6. The f.e.t. is also only allowed to contribute a small amount of the total resistance between pin 2 of the i.c. and common, and this fraction is determined by R_3 in the source lead. Linearizing produces distortion better than 0.05% compared with around 0.4% without.

Transistors Tr_2 and Tr_3 form a long tailed pair which compares the oscillator amplitude with a direct reference voltage. Resistor 8 prevents loading of the oscillator output by changes in Tr_2 input impedance over each cycle. The direct error voltage feeds the f.e.t. gate after filtering (R_7, C_{10}) to remove oscillator components. The two transistors are identical types and mounted together so that their two base-emitter junctions provide temperature compensation; the use of a matched pair in a single can does not seem to be justified. Stability of the d.c. reference is assured by using low temperature coefficient resistors for R_{11} and R_{12} as well as a stabilized negative line.

Though a square wave oscillator followed by a filter could have produced similar amplitude stability simply by defining the voltage excursion of the square wave generator there is a unique advantage in the method described. Namely, the long-tailed pair comparator need not look at the oscillator output directly; it could look at the level of 38kHz which feeds into the multiplier and thus act as a servo, taking up gain drift in the doubler, amplifier and notch filter.

Printed boards (a total of three) are available for this encoder for £7.50 inclusive from M. R. Sagin, 23 Keyes Road, London NW2.



Appendix

Inherent crosstalk arising from deficiencies in the coded signal.
Crosstalk is expressed relative to the full level on the decoded channels when $A=B=1$ as this is the most convenient reference when making measurements.

Amplitude error between the M and S signals.
Ignoring the pilot signal and considering an error δ so that the composite signal becomes

$$\frac{A+B}{2}+\frac{A-B}{2+\delta}\sin 2f_ct$$

i.e. S is low in level if δ is positive.
After multiplication in the decoder, considering only the 38kHz component of the reinserted carrier waveform

$$D(t)=\left|\frac{A+B}{2}+\frac{A-B}{2+\delta}\sin 2\pi f_ct\right|\sin 2\pi f_ct$$

Adding $\frac{1}{4}(A+B)$ to give the decoded A signal, and considering only baseband components gives

$$\frac{A+B}{4}+\frac{A-B}{2(2+\delta)}.$$

Related to peak level, $\frac{1}{2}B$, fractional crosstalk is $\delta/2(2+\delta)$.

Phase error between M and S signals.
Suppose $A=0$, $B=1$ and $B(t)=\sin 2\pi f_Bt$. If a delay of δt exists on the S signal, the composite signal is

$$\frac{\sin 2\pi f_Bt}{2}-\left(\frac{\sin 2\pi f_B(t+\delta t)}{2}\right)\sin 2\pi f_ct.$$

After decoding, adding $\frac{1}{4}(A+B)$, i.e. $\frac{1}{4}\sin 2\pi f_Bt$, and neglecting non-baseband components, the decoded A signal $\frac{1}{4}\sqrt{(1-\cos 2\pi f_B\delta t)^2+\sin^2 2\pi f_B\delta t}$ $=\frac{1}{2}\sin\pi f_B\delta t$, so that fractional crosstalk is $\sin\pi f_B\delta t$.

Error in pilot phase.
Suppose pilot is $0.1\sin f_c(t+\delta t)$, so in the decoder the regenerated 38kHz is $\sin 2\pi f_c(t+\delta t)$. $D(t)=$

$$\left|\frac{A+B}{2}+\frac{A-B}{2}\sin 2\pi f_ct\right|\sin 2\pi f_c(t+\delta t).$$

Add $\frac{1}{4}(A+B)$ and neglecting non-baseband terms the decoded A signal is

$$\frac{A}{4}(1+\cos 2\pi f_c\delta t)+\frac{B}{4}(1-\cos 2\pi f_c\delta t).$$

and fractional crosstalk is $\sin^2\pi f_c\delta t$.

Amplitude imbalance between the sidebands of the S signal.
If $A=0$, $B=1$ and $B(t)=\sin 2\pi f_Bt$ and a sideband imbalance exists then the composite signal is

$$\frac{\sin 2\pi f_Bt}{2}-\frac{\cos 2\pi(f_c-f_B)t}{4}+\frac{\cos 2\pi(f_c+f_B)t}{4+\delta} \qquad \text{A1}$$

Considering only baseband terms

$$D(t)=-\sin 2\pi f_Bt.\left|\frac{8+\delta}{8(4+\delta)}\right|.$$

Adding $\frac{1}{4}(A+B)$ the decoded A signal is

$$\sin 2\pi f_Bt\left|\frac{\delta}{8(4+\delta)}\right|.$$

and fractional crosstalk is therefore

$$\frac{\delta}{4(4+\delta)}.$$

Phase shift in the upper sideband of the S signal.
Taking equation A1 but for a phase error in the (f_c+f_B) component, signal is

$$\frac{\sin 2\pi f_Bt}{2}-\frac{\cos 2\pi(f_c-f_B)t}{4}+\frac{\cos 2\pi(f_c+f_B)(t+\delta t)}{4}.$$

$$D(t)=\tfrac{1}{8}\left|\sin 2\pi f_Bt.\cos 2\pi(f_c+f_B)\delta t+\cos 2\pi f_Bt.\sin 2\pi(f_c+f_B)\delta t\right|-\frac{\sin 2\pi f_Bt}{8},$$

and the decoded A signal is $\frac{1}{4}\sin(f_c+f_B)\delta t$, giving a fractional crosstalk of $\frac{1}{2}\sin(f_c+f_B)\delta t$.

References

1. Z. Mack, Stereo service generator (in German). Circuit of a switching type coder is given. *Funk-Technik* 1968 p.532.
2. Z. Mack, Comparison of transformerless ring-modulators and cross modulators. *Radio and Electronic Engineer* vol. 44 1974 p.407.

Broadcast stereo coder

2 — Circuit description and construction

by Trevor Brook *Surrey Electronics*

The complete coder is shown in Fig. 10. IC_1 and IC_2 provide regulated and short-circuit protected plus and minus 15-volt lines. The output voltage of these i.cs has reasonable temperature stability, which is desirable for the negative line, since it provides the reference for oscillator amplitude. Though short-circuit protected, the regulators cannot withstand reverse polarity at their outputs, so D_{16} and D_{17} prevent damage, should the two supplies be inadvertently shorted together.

The 19 kHz sine-wave oscillator described in part 1, IC_3, has one addition, the chain of diodes D_{11-14} across the output. There is the chance that, when starting, the oscillator output could hit the supply rails and thus go beyond the linear region of the multiplier, IC_4. When the multiplier is overdriven its output, instead of rising further, distorts and begins to fall, which means that the comparator no longer receives an input in proportion to the oscillator amplitude and the oscillator stays locked into a condition where it oscillates at the supply clipping point. Diodes 11 to 14 clip the oscillations below the multiplier's serious non-linearity level without affecting the oscillator distortion when running normally, at the designed output of 1 volt r.m.s.

Multiplier IC_4 has its X+ and Y+ inputs tied together, so that it acts as a linear frequency doubler with R_{23} providing trimming of 19 kHz feed-through rejection. The rejection figure obtainable worsens as the multiplier's maximum permissible input swing is approached, hence the reason for driving at 1 volt.

The loss occurring in the multiplier is recovered by IC_5 and, since it must provide over 30dB gain, a wide bandwidth op-amp is used, a 531. A 748 can just about manage the job but it introduces a significant temperature-dependent phase shift, a very undesirable characteristic in this part of the circuit.

Notch filter IC_6 has virtually unity gain at 38 kHz and is within the capabilities of a 748. Of all the active notch arrangements I have tried, the Wien bridge seems the most repeatable. No very high impedances are involved, the loss at double notch frequency is less than 0.2dB, the corresponding phase shift is small and stable, and a notch deeper than 30dB can be obtained at 19 kHz. Two adjustables set the time constant of one bridge arm and the circuit Q and both are adjusted for the deepest notch. Perhaps IC_6 and its associated circuitry is a lot of trouble to avoid a simple LC rejector; but custom-wound inductors are also a lot of trouble, have poor tolerance and the possibility of causing distortion if ferrite cored.

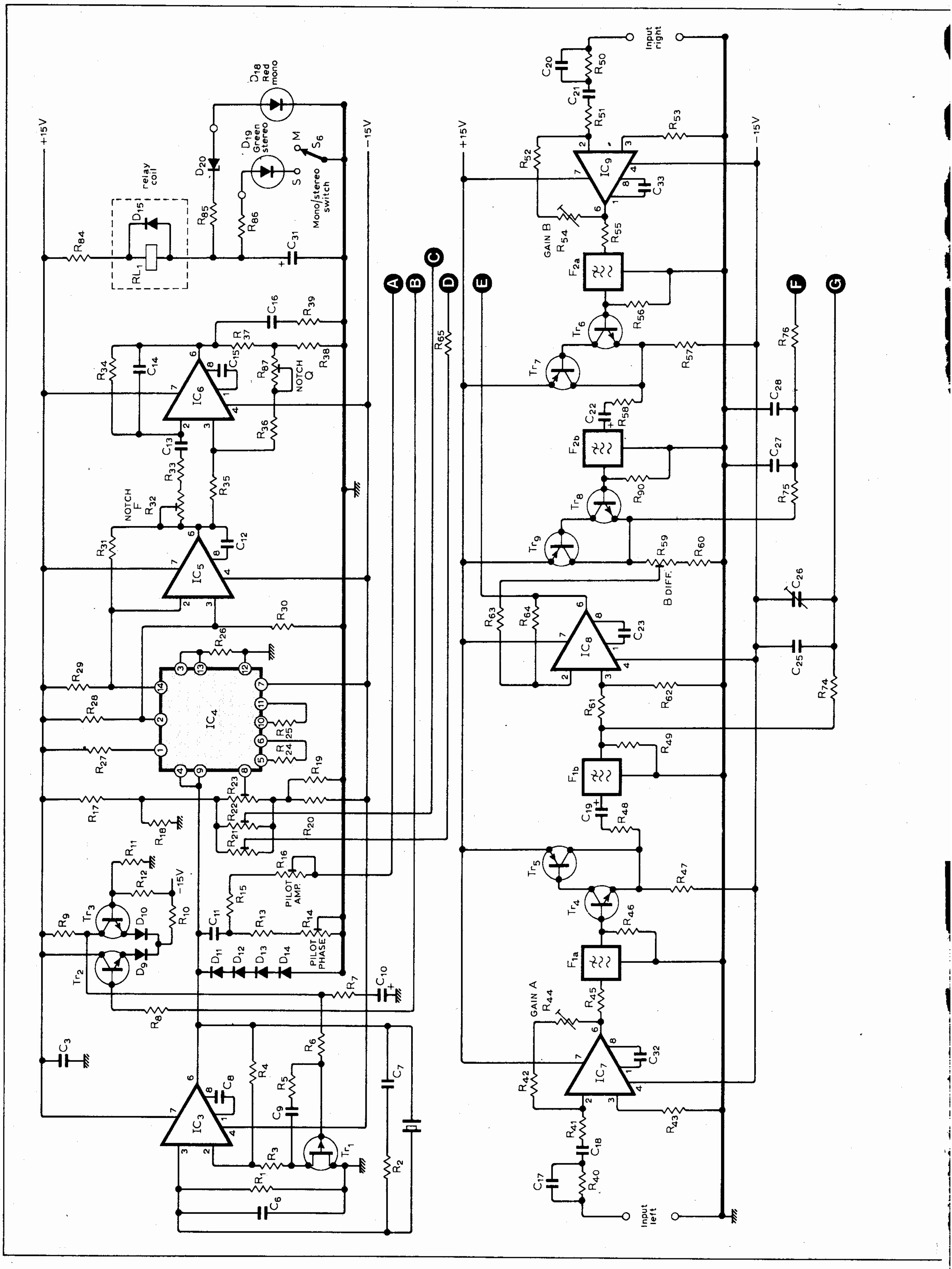
Capacitor 16 couples the 38 kHz into the balanced modulator and blocks the accumulated d.c. offset. Though only a volt or so, it is unlikely to be temperature stable so R_{39} establishes a stiff grounding for the multiplier. The value of C_{16} is chosen with R_{39} to cause small phase shift, yet provide some welcome roll off at low frequencies, since the 531 is a disgustingly noisy little animal. The comparator sensing point is also taken from here, again with no worries about superimposed d.c.

Left-channel audio passes through R_{40} and C_{17} where it receives pre-emphasis of 50μs. Capacitor 17 may be omitted for a flat frequency response or a link could replace R_{40} on the board and R_{40} be placed by a switch bank with various capacitors to give a choice of pre-emphasis. A straightforward audio amplifier IC_7 drive the first filter section through its correct source impedance, R_{45}. The filter is terminated by R_{46} and feeds into a compound emitter follower, Tr_4, Tr_5: single-transistor emitter followers cause too much distortion, even at signal levels below 1 volt as here. Resistor 48 is the source impedance for F_{1B} which is terminated by R_{49}. Arrangements on the right channel are identical apart from F_{2B}'s terminating resistor which is split between a preset, R_{59}, and a fixed resistor. These filters are normally intended for use as a stereo pair, but on an experimental coder there appeared a surprisingly large phase shift between the M and S signals as 15 kHz was approached. This turned out to be due to crosstalk (at −60dB) between the two halves of the filter which produced a spurious signal of different phase on the 'silent' channel. The cure adopted here is to feed each channel back through the second half of its original filter block and keep the left and right channel blocks well apart.

The A and B signals emerging from F_{1B} and F_{2B} are fed via their phase shifting networks, R_{74}, C_{25}, C_{26} and R_{75}, C_{27}, C_{28}, to the output adder IC_{11}. The different values for C_{25} and C_{27} is explained by different paths through the differencing amplifier and difference in circuit board capacity for the two channels.

The differencing amplifier, IC_8, uses a 748 rather than a 741, since less phase shift is introduced at the higher audio frequencies and the change with temperature of the remaining phase shift is lower. The second drawback of the multipliers used here is that they produce a small amount of second harmonic distortion and, though this is immaterial in the doubler configuration, it is relevant when using the balanced modulator configuration. Such distortion on the audio port will produce second harmonic distortion for difference signals below 7.5 kHz and beat tone distortion for frequencies between 7.5 and 15 kHz. On the 38 kHz port, the effect will be to give an output, with associated sidebands, at 76 kHz. Like feedthrough, these effects worsen as the multiplier is driven harder and here the carrier level, and audio level for a full difference signal, are set 6dB below the multiplier's non-linearity point. The audio takes precedence and goes to the X port, which has the better linearity specification. The objection to driving the balanced modulator at even lower levels is that noise would become obtrusive. The double-sideband, suppressed-carrier difference signal from IC_{10} is fed to the adder at the correct level via R_{72}.

The gain of 15dB required from IC_{11}, the output adder, for the S signal, is possible from a 748 and the noise level of these devices is also good enough for this position. The signal components may be switched individually by the d.i.l. switch mounted on the board, S_{1-4},


+15V
−15V
relay coil
Red mono
Green stereo
Mono/stereo switch
NOTCH F
NOTCH Q
PILOT AMP.
PILOT PHASE
GAIN A
GAIN B
B DIFF.
Input right
Input left
IC3
IC4
IC5
IC6
IC7
IC8
IC9

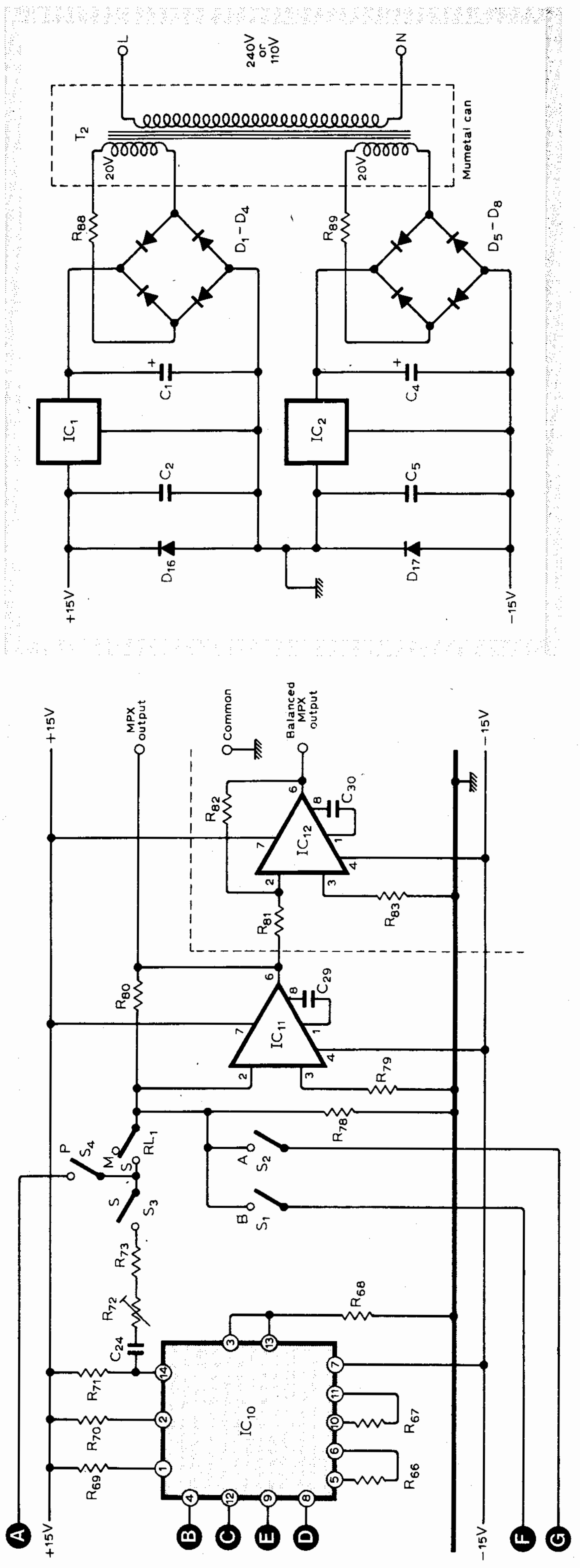


Fig. 10. Circuit diagram of complete coder. (The junction of C_{16} and R_{39} should be taken to be the line marked B.)

and R_{78} is present to stop the 748 going unstable should all the switches be turned off. A balanced output is produced by IC_{12}, which is a unity gain inverter.

The sine-wave pilot signal is taken directly from the oscillator output, passed through a trimmable phase shift network C_{11}, R_{13}, R_{14}, and then attenuated suitably by R_{15}, R_{16} before reaching the adder. Mono/Stereo switching is achieved by a reed relay mounted on the board immediately by the adder, which disconnects the pilot and S signal. The reverse diode and capacitor around the relay coil completely remove any click due to the switch but some click remains as the reed contacts make or break. There is no d.c. offset being switched and no capacitor charging as the contacts close and the click only occurs if the pilot is switched on at the d.i.l. switch. The reason is that the 19 kHz sine wave is being interrupted instantaneously: another way of thinking of it is 100% amplitude modulation, and a continuum of sideband energy will extend from d.c. to infinity. The peak level of the click at the coder output viewed on a 'scope with a 15 kHz filter and no de-emphasis is −30dB. Some coders leave the S signal on when in the mono mode but it is no trouble here to remove it and it seems good practice to do so if stereo performance is not compromised.

The little arrangement around the red and green l.e.d.s allows mono and stereo indicators to operate along with the reed relay, while only using a single-pole switch contact, which closes for stereo. This allows for easy remote switching. The green l.e.d. passes full relay current in stereo and the red l.e.d. draws a small current in mono which is insufficient to hold the relay in, yet subjectively gives the same brightness because of the greater efficiency of red l.e.d.s.

Construction

To achieve a compact layout, as well as to avoid links and keep signal tracks short in the interests of reducing crosstalk within the coder, the p.c. board has to be double sided. The whole coder, including its power supplies and mains transformer, is accommodated on a board 165 by 165mm. To avoid hum pickup it is essential for the board-mounted mains transformer to be magnetically shielded. Though the board track layout is designed to avoid ground loops, many i.cs have built-in loops which make them susceptible to hum induction when in a magnetic field in the same plane as the i.c. chip. This applies particularly to the multipliers and regulators used here, and a cylindrical Mumetal can for the mains transformer provides over 30dB reduction in its hum field, a more than adequate margin. The heatsinks provided for the regulators run hardly warm to the touch and only reach 30°C above ambient under supply overload conditions. However, their sides provide convenient points for glueing down the large smoothing capacitors to prevent them from vibrating and their leads fracturing under severe mechanical shock. Clear Bostik 1 is suitable for the purpose.

All the trimmers are visible-setting, single-rotation types. None of them is doing more than providing a very fine trimming adjustment, so multiturn types are not justified. In addition, being able to see the position of a preset is extremely useful as an unusual setting frequently leads to discovery of an incipient fault.

Resistors which have a bearing on gain, phase or important time constants are 2%, with thick film types being preferred for the lower values where they are available, since they have a lower temperature coefficient (±100ppm) than the 2% metal oxide types (±250ppm). Similar comments apply to capacitors where 1% silver mica types are used for the notch filter and pilot phase corrector with a low temperature-coefficient polycarbonate type for C_{16}. Stripboard construction is not likely to be successful, but printed circuit boards are available from the address at the end. Ground tracks radiate along the board from the output adder and there are in addition several, apparently redundant, ground tracks forming ground guards to reduce board leakage and intertrack capacity. The positioning of circuit sections on the board also contributes to minimal 19 or 38 kHz pickup along the audio paths or by the output amplifier. The long-tail pair comparator transistors in the oscillator are mounted together and a drop of glue between them will do no harm. While the difference signal is at a fairly high impedance, the capacity of its line has to be kept low to avoid loss or phase shift of the upper sideband and this is done by IC_{10} being directly next to the adder.

The board pins connecting the plus and minus 15V lines through the board to their distribution tracks across the top can be omitted until correct functioning of the power supplies has been checked. To simplify initial checking it is a good idea to omit the pre-emphasis capacitors as well, C_{17}, C_{20}, so the coder can be set up with a flat frequency response.

Printed circuit boards
A set of p.c.bs comprising one double-sided board, which measures 6½ × 6½in, and two smaller single-sided boards is available at £7.50 inclusive from M. R. Sagin at 23 Keyes Road, London N.W.2.

X_1 19kHz crystal, RC 13U (Surrey Electronics, The Forge, Lucks Green, Cranleigh, Surrey).

Mains transformer (Surrey Electronics).

F_1, F_2. BLR2011N filters (Harrogate Radio Ltd, 2/3 Sykes Grove, Harrogate, W. Yorks).

Heat sinks. Redpoint TV3 (Elecrovalue, 26 St. Jude's Road, Englefield Green, Egham, Surrey).

Relay, d.i.l. switch, trimmers and trimmer capacitors can be obtained from Doram Electronics, PO Box TR8, Wellington Road, Industrial Estate, Wellington Bridge, Leeds 12.

The next article will describe the alignment of the decoder.

Parts list

R_1	1.8k
R_2	1.8k
R_3	18k
R_4	39k
R_5	1M
R_6	1M
R_7	470
R_8	1k±1%
R_9	1M
R_{10}	47k
R_{11}	1.8k±2%
R_{12}	8.2k±2%
R_{13}	8.2k±2%
R_{14}	2.2k
R_{15}	330k±2%
R_{16}	100k
R_{17}	10k±2%
R_{18}	470±2%
R_{19}	470±2%
R_{20}	10k±2%
R_{21}	4.7k
R_{22}	4.7k
R_{23}	4.7k
R_{24}	8.2k±2%
R_{25}	8.2k±2%
R_{26}	5.6k±2%
R_{27}	3.3k
R_{28}	3.3k
R_{29}	3.3k
R_{30}	100k±2%
R_{31}	150k±2%
R_{32}	470
R_{33}	1.5k±2%
R_{34}	1.8k±2%
R_{35}	47k±2%
R_{36}	18k±2%
R_{37}	1.8k±2%
R_{38}	47k±2%
R_{39}	1ok±2%
R_{40}	100k±1%
R_{41}	6.8k±2%
R_{42}	39k±2%
R_{43}	39k
R_{44}	22k
R_{45}	1k±2%
R_{46}	4.7k±2%
R_{47}	3.3k
R_{48}	1k±2%
R_{49}	4.7±2%
R_{50}	100k±2%

R_{51}	6.8k±2%
R_{52}	39k±2%
R_{53}	39k
R_{54}	22k
R_{55}	1k±2%
R_{56}	4.7k±2%
R_{57}	3.3k
R_{58}	1k±2%
R_{59}	470
R_{60}	4.3k±2%
R_{61}	100k±2%
R_{62}	470k±2%
R_{63}	100k±2%
R_{64}	470k±2%
R_{65}	8.2k
R_{66}	8.2k±2%
R_{67}	8.2k±2%
R_{68}	5.6k±2%
R_{69}	3.3k
R_{70}	3.3k
R_{71}	3.3k±2%
R_{72}	2.2k
R_{73}	9.1k±2%
R_{74}	4.7k±2%
R_{75}	4.7k±2%
R_{76}	22k±2%
R_{77}	22k±2%
R_{78}	10k
R_{79}	6.8k
R_{80}	47k±2%
R_{81}	6.8k±2%
R_{82}	6.8k±2%
R_{83}	6.8k
R_{84}	470
R_{85}	3.3k
R_{86}	470
R_{87}	4.7k
R_{88}	15 1/2W
R_{89}	15 1/2W
R_{90}	4.7k±2%
C_1	2200µ/40V
C_2	100n
C_3	100n
C_4	2200µ 40V
C_5	100n
C_6	47n
C_7	4.7n±1%
C_8	3.3p
C_9	100n

C_{10}	33µ 10V
C_{11}	1n ±1%
C_{12}	3.3p
C_{13}	4.7n ±1%
C_{14}	4.7n ±1%
C_{15}	3.3p
C_{16}	10n ±5%
C_{17}	500p ±1%
C_{18}	1µ ±5%
C_{19}	6.8µ
C_{20}	500p ±1%
C_{21}	1µ ±5%
C_{22}	6.8µ
C_{23}	3.3p
C_{24}	1µ ±5%
C_{25}	10p
C_{26}	20p
C_{27}	47p
C_{28}	20p
C_{29}	3.3p
C_{30}	3.3p
C_{31}	6.8µ
C_{32}	5p
C_{33}	5p
D_1	1N4001
D_9 – 14	1N914
D_{15}	1N4001
$D_{16, 17}$	1N4001
D_{18}	Red l.e.d.
D_{19}	Green l.e.d.
D_{20}	6.2V Zener
Tr_1	2N5457
Tr_{2-4}	BC239C
Tr_5	BC309
Tr_6	BC239C
Tr_7	BC309
Tr_8	BC239C
Tr_9	BC309
$IC_{1, 2}$	L131 or TDA1415
IC_3	748
IC_4	MC1495L
IC_5	531
$IC_{6, 7, 8, 9}$	748
IC_{10}	MC1595L

Broadcast stereo coder

3—Setting up

By Trevor Brook, *Surrey Electronics*

A practical design for a high quality coder suitable as a test instrument was described in the April & June issues. Apart from the audio filtering, inductors have been avoided and a compact board layout produced. A v.h.f. unit, for servicing checks on receiver performance, could also be used by demonstration showrooms to feed programmes of their own choice to stereo tuners.

Part 1 examined the stereo multiplex system and established tolerance limits for signal components. Channel separation was considered as this would assume increased importance if a matrix system of surround sound broadcasting were adopted. Part 2 gave construction details and alignment details follow in this part. Part 4 gives modifications to the Portus and Haywood decoder to provide a low distortion reference decoder.

In this setting-up procedure 0dB level refers to 0.775V.

—With the coder in mono, set A and B gains, by means of R_{44}, R_{54}, so that 0dB input at 1kHz gives —7dB at the output. Check that the amplitude response is +0.5, —1.0dB from 20Hz to 15kHz relative to the 1kHz level for each channel.

For measurements near 15kHz, a frequency counter will prove most useful if the audio signal generator calibrations are not accurate.

—With a grounded crocodile lead on the "oscillator defeat" pin, IC_3 pin 6, check the distortion for each channel with 0dB output at 1kHz. A reading of better than 0.03% will confirm that all is well.

—To align the 38kHz path, set presets R_{23}, R_{32} and R_{87} to mid-position and remove the "oscillator defeat" link.

—Looking at the output of IC_5, pin 6 on an oscilloscope, adjust R_{23} for a rough null in 19kHz content on the 38kHz waveform.

—Connect a nulling distortion meter, tunable to 38kHz, to pin 4 of IC_{10}. (Many distortion meters only cover up to 20kHz, but generally they are easily modified by soldering an extra parallel resistor in each arm of the null bridge so that the upper frequency becomes 40kHz. For the job here accuracy is not very important; all that is required is good rejection of the 38kHz so that the remaining 19kHz component can be nulled.) Looking at the distortion meter output on a 'scope adjust R_{32} and R_{87} alternately to achieve the best rejection of 19kHz.

—Final trimming of R_{23} as well should leave no 19kHz visible amongst the noise, and better than 60dB below the 38kHz level.

—The 38kHz amplitude at this same point may now also be checked as +8dB ±0.5dB.

—With the oscillator system now set up properly the distortion of the 19kHz at pin 6 of IC_3 can be checked as below 0.1%.

—Switch the coder to stereo and look at the 38kHz at the output, with only the S switched on at the d.i.l. switch. Adjust for minimum carrier with R_{22}. Using broadband metering the 38kHz null will be masked by the residual 76kHz generated by IC_{10}, which does not null out.

—Feed 1kHz at around 0 to +6dB into the left channel and defeat the oscillator. Still with only S switched on adjust R_{23} for a null of audio leak through in IC_{10}.

—Allow the oscillator to run and feed 1kHz at 0dB into the left channel with A, B and S turned on at the d.i.l. switch. Lock the 'scope to the audio and adjust R_{72} for the roughly correct M/S amplitude relationship seen in Fig. 11(a).

—Repeat for the right channel input but this time leave R_{72} alone and adjust the B difference pot, R_{59}.

—Switch the pilot on at the d.i.l. switch and, with no audio input, set its level to —21dB at the coder output, using R_{16}.

—Feeding 1kHz at around —10dB into either left or right channels, turn on only the S and pilot at the d.i.l. switch. Locking the 'scope to the audio should display an "eye" pattern, as in Fig. 11(b). The correct pilot phase is when the eye appears symmetrical and this is more easily seen with some vertical magnification arranged as shown in Fig. 11(d). Resistor R_{14} adjusts the pilot phase and the effect of a slightly incorrect setting is seen in Fig. 11(e).

Table 2: Measurements on prototype coder

No pre-emphasis	
Frequency response +0.5 dB, —1.0 dB	20Hz to 15KHz
Rejection of 19kHz	68dB
Rejection of frequencies above 19kHz	58dB
Crosstalk at 20°C 20Hz-15kHz	55dB
Crosstalk 10-40°C 20Hz-15kHz	45dB
Residual 38kHz	50dB
Pilot phase accuracy	1°
Beat tone distortion, 15kHz full M, full S or L or R overdriven 6dB	0.1%
Spurious responses above 53kHz, full M, full S or L or R overdriven 6dB:	
sidebands of 57kHz	—63dB
carrier and sidebands at 76kHz	—48dB
carrier and sidebands at 152kHz	—84dB
Measurements using reference decoder (part 4) and 50μs de-emphasis:	
harmonic distortion, 1kHz full M, S, L or R	0.04%
signal-to-noise ratio, 20Hz to 15kHz, mean reading meter, unweighted:	
mono	79dB
stereo	71dB

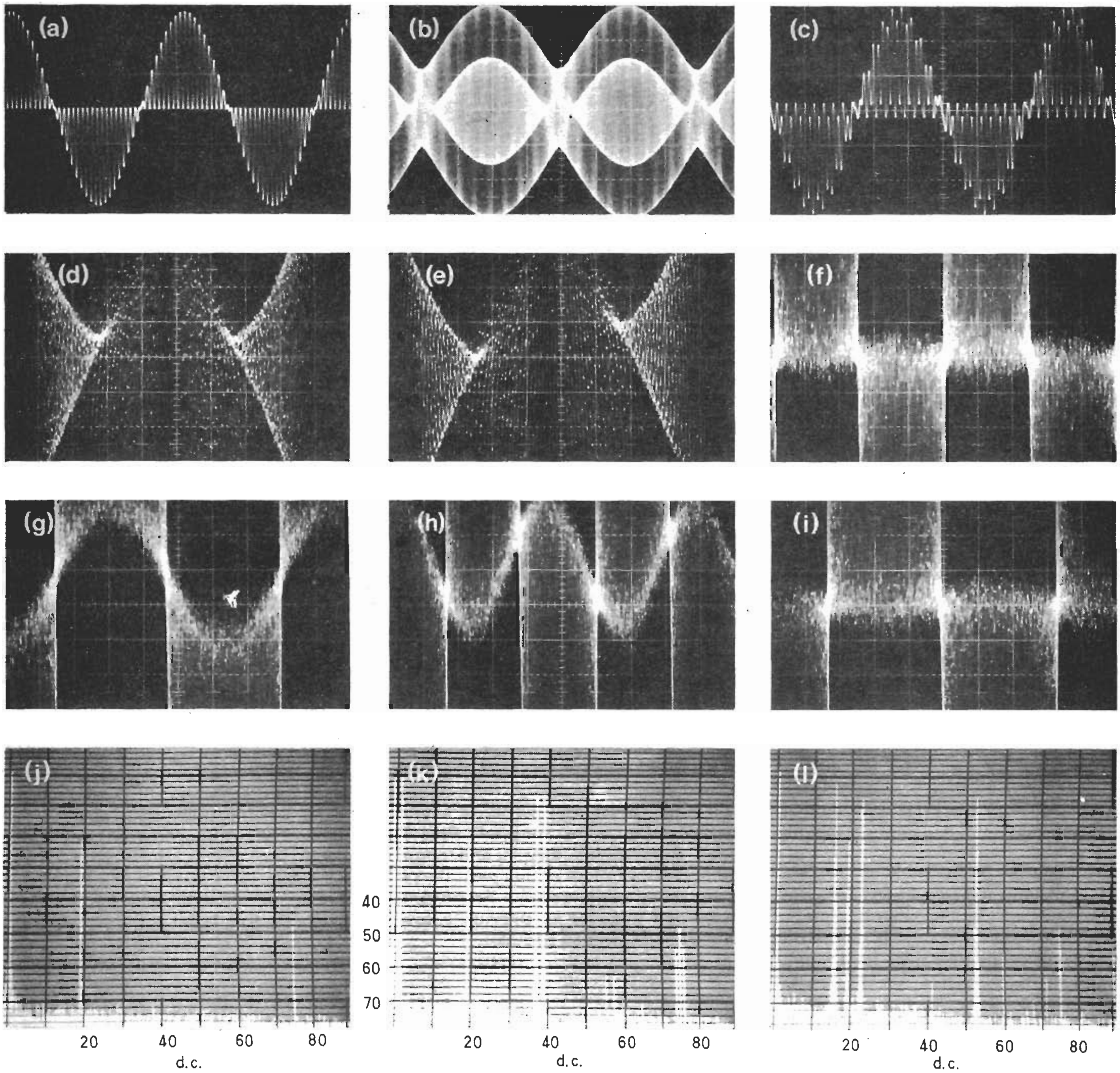


Fig. 11. Correctly set-up coder, with full 1kHz A signal and no pilot is seen at (a) which indicates the flat zero line. The pilot phase "eye" is at (b), with only S and pilot, while (d) shows the zero crossing region of (b) magnified virtually by a factor of 100 – (e) is the same but with an incorrect pilot phase setting. Zero line ripple at (f), with full 1kHz A signal, is obtained by X100 vertical gain and clipping amplifier shown in Fig. 12. Result on zero line ripple of low S amplitude is at (g), while (h) shows too high an S amplitude in wrong phase. Photo (c) is of composite multiplex signal with pilot and full 1kHz signal. Zero-line ripple at (i) is that obtained for 15kHz with coder correctly set up. Spectrum analyser photo at (j) shows noise spectrum when in stereo mode but with no audio. Pilot at 19kHz and main spurious response – 76kHz at –48dB can be seen. Stereo spectrum with A overdriven by 6dB with 1kHz is at (k) and with 15kHz at (l). Analyser measurements were performed by Marconi Instruments TF 2370, 50Hz bandwidth, direct into 50Ω input via 3.3kΩ resistor, not using high-impedance probe.

—Pilot amplitude and phase adjustments are very slightly interdependent, so repeat the last two adjustments.

Clipping amplifier

If all is well to this point, then channel separation will exceed 40dB at 1kHz, but to see the M/S amplitude and phase error more easily for greater separations requires × 100 vertical magnification compared with that in Fig. 11(a). Some 'scopes may manage this without overloading, but most do not so a useful amplifier and clipper circuit is given in Fig. 12. This is simply a 20dB amplifier with diodes arranged to bring the gain below unity as soon as the output swing exceeds 0.6 volts. The amplifier has quick recovery from the clipping so does not degrade the interesting zero voltage area of the stereo waveform.

The clipping amplifier can conveniently be built on a scrap of Veroboard and placed inside a metal 35mm film can. The output resistor stops r.f. instability when driving capacitive loads in the clipping condition. Used in conjunction with a directly-coupled 'scope giving 20dB gain (which should not cause overloads) and at least 5MHz bandwidth, the required × 100 magnification with low phase shift is achieved and a correct stereo waveform appears in Fig. 11(f).

—Using the clipper arrangement repeat first two items in column 3, page 89.

The only limitation to correct setting should be the noise along the zero line of the waveform. Figure 11(g) shows the in-phase zero-line ripple caused by low S signal amplitude corresponding to a loss in S of 2.7%.

—Now change the input frequency to 15kHz and adjust the M/S phase accuracy (C_{26} adjusts for the A and C_{28} for the B channel).

Phase errors appear on the 'scope

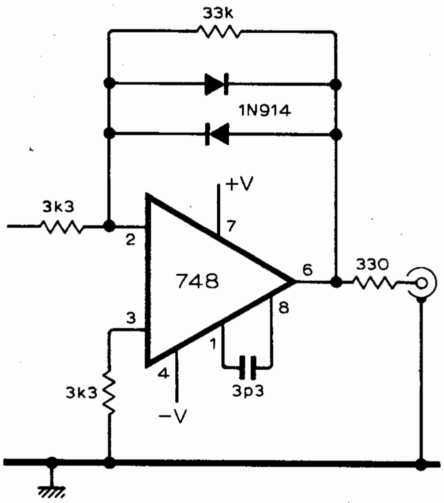


Fig. 12. Amplifier with 20dB gain and clipping arrangement to allow large vertical magnifications of the zero-voltage region without overloading oscilloscope Y amplifiers.

display as a sine-wave zero-line ripple shifted in phase relative to the main pattern and Fig. 11(h) shows the appearance of amplitude and phase errors combined.

—Check now that M/S accuracy is maintained over the whole audio frequency range.

If an audio signal in antiphase and of precisely the same level is available the initial adjustment can be improved upon.

—Feed antiphase audio at 1kHz into left and right channels at 0dB and with the A and B signals only switched on make a very slight trimming adjustment to either R_{44} or R_{54} so that the output nulls. This gives a more accurate channel balance than setting channel gains up on a millivoltmeter.

The audio leakthrough set on p.89 with the oscillator stopped can be adjusted under working conditions for the multiplier if a 15kHz low-pass filter is available.

—Connect to the coder output and with only the S signal turned on at the d.i.l. switch feed left or right input with 15kHz at +6dB (or feed both left and right with antiphase both at 0dB). Potentiometer R_{23} is adjusted for a null in 15kHz leakthrough. Correct setting of R_{23} is important, otherwise false settings in R_{72}, C_{26} and C_{28} can be produced.

Without a spectrum analyser or decoder an estimate of the beat-tone distortion may be made with the aid of a distortion meter.

—Continue as above for the audio leakage check but switch on A, B and the pilot as well as S. Use the distortion meter to null the 15kHz and some of the beat tones will give a reading below 0.1%. (This reading is only an indication that all may be well as no account has been taken of beat tones above 15kHz which will be heterodyned into the audio range in decoding. If a frequency counter is available a check can be made that the pilot frequency is 19kHz ±2Hz.)

—Check temperature stability by feeding 1kHz at 0dB into the left channel, with A, B and S turned on at the d.i.l. switch. Lock the 'scope to the audio and with the × 100 vertical magnification arrangement view the change in relative M/S amplitude and phase. With a temperature rise from 20 to 40°C the S amplitude should fall by 0.8%, i.e. 24mm on a pattern magnified to 6000mm.

Allow at least half an hour for all components on the board to reach the new ambient temperature. What S amplitude loss there is can be shown to be predominantly due to the balanced modulator i.c. by briefly holding a soldering iron on its case. With the methods described here no phase error between the M and S components should be visible over the whole temperature range +10 to +45°C. Incidentally it is quite impossible to align a coder for channel separation by using a decoder, as apparently good separation can be achieved on a particular decoder with quite the wrong phase and amplitude settings.

Three checks on performance can be made using a suitable reference decoder, such as the modified Portus and Haywood design described in Part 4. Using 50μs de-emphasis the noise level referred to 1kHz full level (−1dB at the coder output) should be ⩾−70dB, unweighted, mean reading meter, 20Hz to 15kHz. Again with de-emphasis, readings of coder-decoder harmonic distortion for 1kHz full A, B, M or S should be 0.04% and the 15kHz beat tone under the same conditions 0.35%.

Some of the distortion above is contributed by the decoder and the only satisfactory way of assessing the purity of the coder output is by spectrum analysis. Figure 11(j) shows the coder noise spectrum when switched into stereo. The 19kHz pilot tone is at −21dB and the slight mark at 38kHz is the suppressed 38kHz carrier at −71dB. The spurious 76kHz double frequency output from the balanced modulator is at −48dB.

Figure 11(k) shows 1kHz in left or right channels overdriven by 6dB. The baseband signal is at −1dB, normally only reached for full M signal, i.e. full A and B in phase. After the pilot are the two S signal sidebands at −7dB, normally only reached for full S signal i.e. full A and B in antiphase. Above this are two spurious responses, sidebands of 57kHz and the 76kHz signal again. For 1kHz, the 57kHz components are harmless, but for higher audio frequencies the lower sideband of the pair falls into the S signal band. On this photo it is also interesting to notice the slight noise modulation effect (about 4dB) which only becomes visible when the S signal is within 3dB or so of full amplitude.

Figure 11(l) shows the situation as above (6dB left or right overdrive) but with 15kHz input. Apart from the lower sideband of 57kHz, 42kHz at −64dB, other minor beat tones are visible at 4kHz, −67dB, and 7kHz, −68dB. The line at 27kHz seems to have been a noise peak, since it bears no obvious arithmetical relationship with the frequencies involved and does not appear in other photographs taken at the time. The 42kHz component will demodulate to 4kHz at −63dB in the left and right channels and this would indicate a beat tone figure for the coder of 0.1%, and with 50μs de-emphasis 0.07%.

Not covered on the photographs, the only component observed above 100kHz was 152kHz and associated sidebands at −84dB. For decoder and receiver measurements the 76kHz outputs are not troublesome — the presence of odd harmonics would have been more worrying — but for some purposes the use of a precision multiplier might be desirable.

To be concluded.

Correction. In the circuit diagram (on page 76, June issue) capacitors C_{25} and C_{26} should have been shown earthed, rather than returned to the −15V rail, and C_{28} shown variable. The G lead should have R_{77} inserted, and the junction of C_{16} and R_{39} should connect to lead B. Resistor R_{75} should be taken to the upper end of R_{59}, and not Tr_8 emitter, which itself should connect to Tr_9 collector through a 33μF capacitor. Emitter of Tr_8 should have R_{91} connecting it to the −15V rail. Capacitor C_{22} should be short-circuited. In the components list R_{49} is 47kΩ, R_{91} is 3.3kΩ, and C_6 is 4.7μF ±1% and not 47μF. Resistor R_8 can be 2%.

Broadcast stereo coder

Three decoders assessed, a reference decoder circuit, filters, and a v.h.f. oscillator

by Trevor Brook, Surrey Electronics

This article concludes the series on the high-quality stereo coder design with a low-distortion decoder circuit. Performance details of the coder, assessed using this decoder, were given in the October issue.

NEED FOR A REFERENCE DECODER for performance checks on the coder prompted an investigation of some commonly available types of decoder. Some decoders produce their best channel separation from a degraded multiplex signal, such as is likely to emerge from the demodulator of present receivers, and the crosstalk measured in Table 3 using an ideal signal is given as a guide to what to expect when testing decoders fed directly from a coder. The setting of the free-running frequency of the phase-locked loop i.c. decoders can also have a considerable effect on channel separation and the best readings obtained are given in Table 3. The 1310 used was the best of seven selected for low mono distortion. All were very similar in stereo but two of the seven gave mono distortion readings of 0.45% on one of their outputs.

The use of a low-pass filter preceeding the decoder is bound to reduce channel separation if it does not have a linear phase characteristic and low amplitude ripple and this effect can be seen in the Skingley and Thompson circuit (*WW* May 1974 page 124). Though a sacrifice in channel separation results, such simple filtering does achieve its purpose of dramatically reducing "birdy" interference from adjacent stations, which otherwise is subjectively far more irritating.

Two odd effects appeared when testing CA3090 decoders using the RCA data sheet circuit. The decoder would trip out of stereo if full level 15kHz M signal was fed into it and limiting of the audio outputs accompanied by large beat tones occurred with full S signal for 15kHz audio. These effects are presumably due to the 15kHz, or lower sideband of the S signal, confusing the 19kHz phase locked loop.

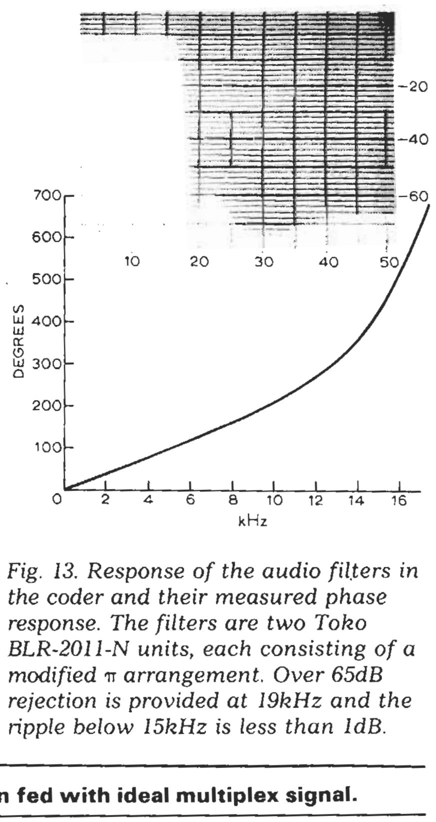


Fig. 13. Response of the audio filters in the coder and their measured phase response. The filters are two Toko BLR-2011-N units, each consisting of a modified π arrangement. Over 65dB rejection is provided at 19kHz and the ripple below 15kHz is less than 1dB.

Table 3. Stereo decoder comparison when fed with ideal multiplex signal.

	Input mV	Distortion (%) mono	stereo 1kHz	stereo 15kHz	Crosstalk dB 1kHz	Crosstalk dB 15kHz
MC1310 CA1310	300	0.09	0.09	0.67	40	37
1310 & filter	300	–	–	–	40	20
CA3090	180	0.17	0.18	1.7 L or R 3 S at –10dB	43	30
Portus & Hayward	600	0.05	0.38	1.3	–	–
P&H modified	600	0.04	0.04	0.35	30	31

Stereo distortion measured at full L, R, M or S level. Worst reading of two channels shown. By altering the pilot phase on the coder channel separation on the modified Portus and Haywood decoder will reach 54dB at 1kHz and 50dB at 15kHz. This has the same effect as adjusting the oscillator trimmers on the 1310 and CA3090 for best channel separation, not necessarily at a free-running frequency of exactly 76kHz.

Finally tested was the Portus and Haywood decoder (*WW* Sept 1970). Needing principally lower harmonic and beat tone distortion, I devised the following modifications, included in the circuit of Fig. 14.

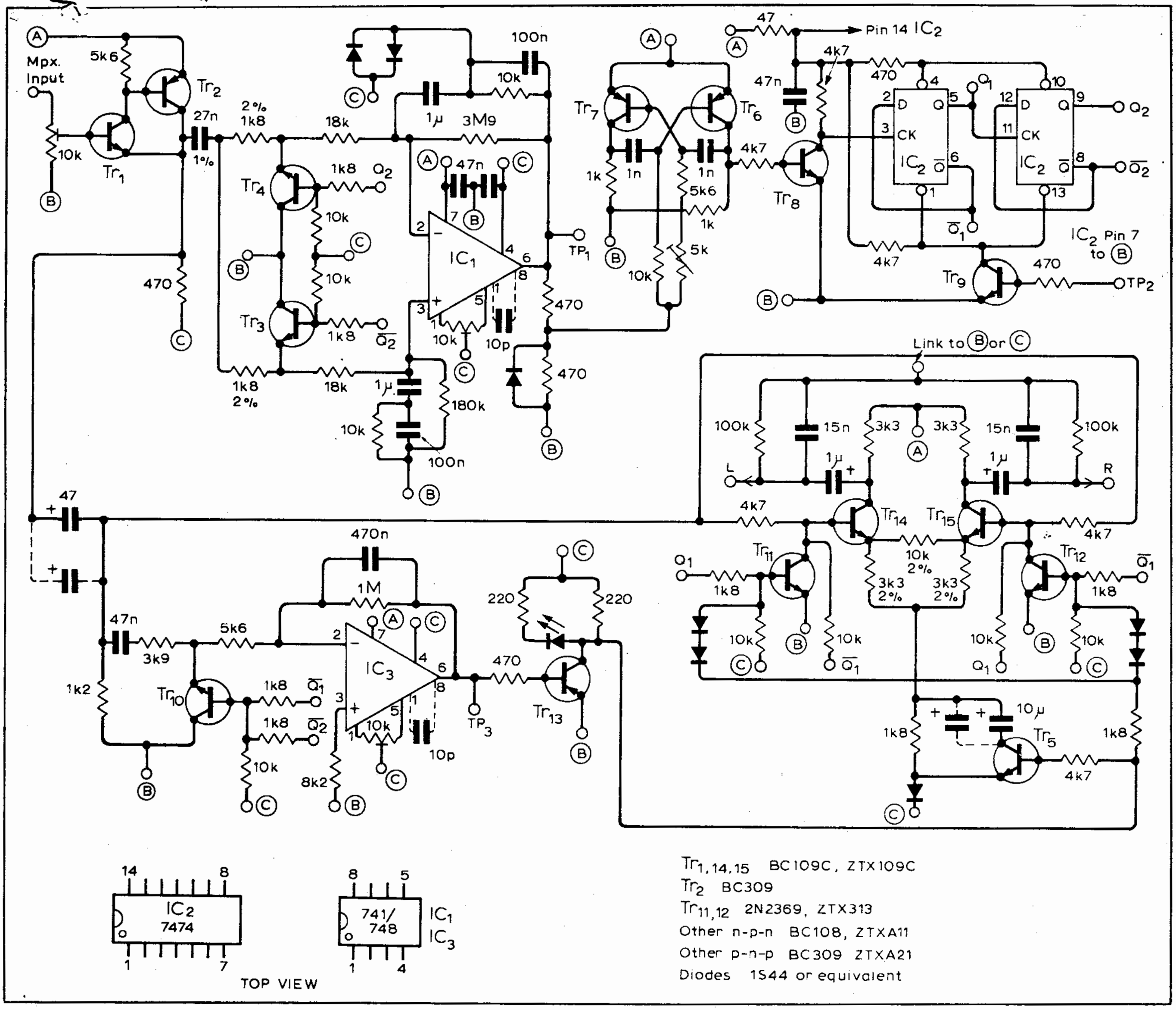
- Change Tr11 and Tr12, formerly BC108 types, for 2N2369, ZTX313 or any high-speed switching transistor.
- Change Tr14 and Tr15 for high-gain audio types, BC109C, ZTX109C, etc.
- Convert the input amplifier to a compound emitter follower, now with a lower emitter resistor and a gain potentiometer at the input. This can be done neatly on the original Integrex p.c. board using only one link. This modification is only suitable if the input amplifier is not required to provide any gain.
- Operate the decoder with only 1.4V at TP2, the pilot level test point, not 1.5V.

These modifications brought the 1kHz distortion in stereo to 0.06% and, with the further suggestion by Mr Portus of fitting pull-up resistors R_{64}, R_{65} onto the bases of Tr14 and Tr15, gives the excellent figures in Table 3 with the only penalties a couple of dB lower audio output and higher switching waveform on the outputs. Low frequency channel separation is easily improved by paralleling 1000μF 10V electrolytics across C_5 and C_{18}. Though irrelevant for normal listening, good separation is desirable when measuring the coder's noise level.

All decoders proved sensitive to supply hum and noise and filtering along the lines shown, Fig. 15, is needed to reduce the noise output from i.c. voltage regulators to allow signal-to-noise measurements beyond 64dB or so.

VHF oscillator

A simple v.h.f. oscillator with a varicap arrangement which has low enough capacity along the multiplex path to avoid h.f. loss is shown in Fig. 16. The oscillator coil is printed on the p.c. board alongside a coupling link which gives roughly 70 ohms output impedance through R_6. Coupling is low en-

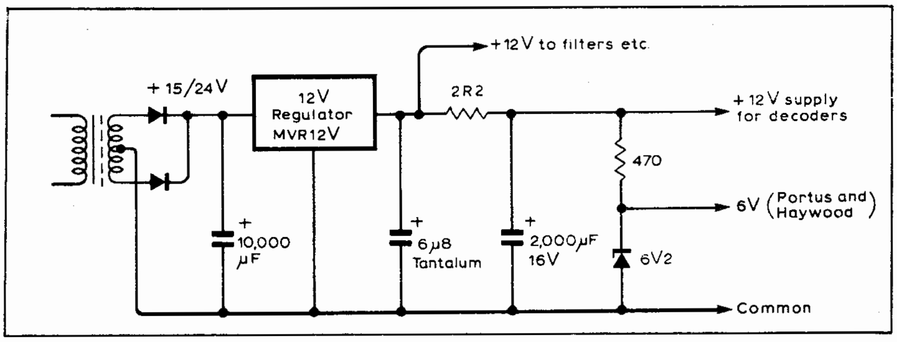


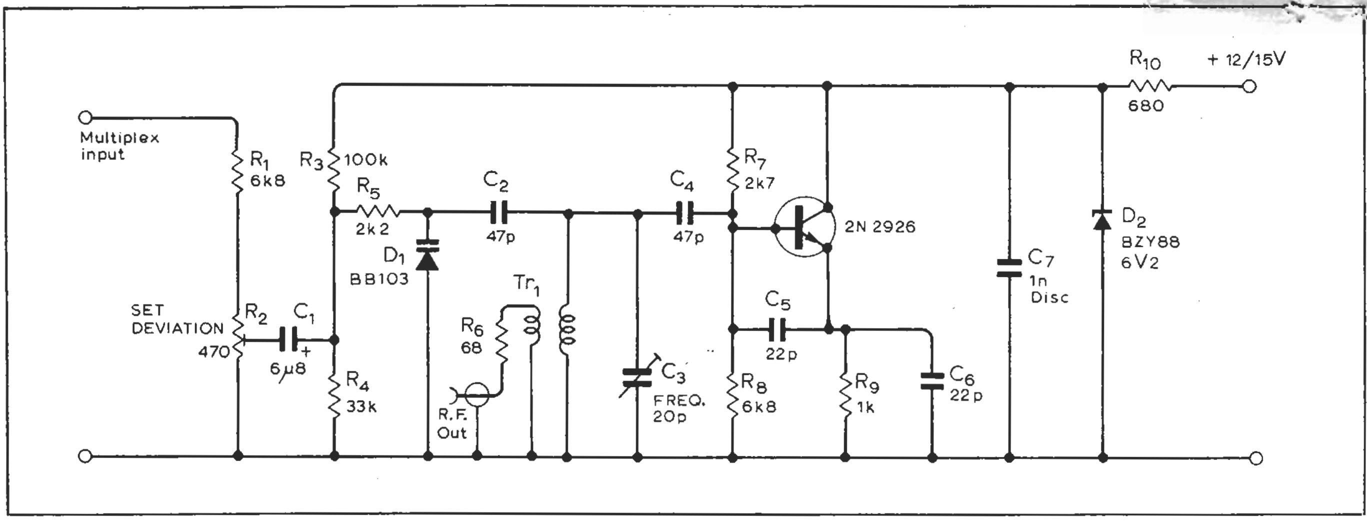
▲ *Fig. 14. Modifications to the Portus and Haywood decoder to improve both distortion and channel separation. Faster switching times and high gain transistors in the matrix with a different input amplifier arrangement give 1kHz distortion better than 0.04%. Voltage levels of points A, B and C can be either +12, +6 and 0V or +6, 0 and −6V respectively.*

ough to avoid frequency jumping with various loads. This device is only intended for use on a fixed frequency and there is no varicap sensitivity or linearity correction. Calculation for this circuit suggests distortion at full deviation of less than 0.5%. For a fully tuneable generator with calibrated attenuator the coder could be fed into the wideband modulation input of the Sound Technology FM1000 signal generator.

On stereo it is important for the deviation to be set correctly. Without an analyser or deviation meter the best way is to measure the pilot tone level before deemphasis when tuned to a BBC stereo station transmitting silence. They tune to the frequency selected for the oscillator and adjust its deviation to produce the same voltage. All the BBC stereo stations I can receive have pilot deviations within 1.5dB of Wrotham Radio 3. The output from the oscillator at around 60mV is adequate to feed a passive distribution system or with coaxial attenuators it can be used for receiver checking. Thirty decibels of attenuation (at 1.9mV) will still keep any reasonable f.m. receiver in full quieting on stereo while a further 6dB attenuation (685mV) will quieten a good tuner.

Fig. 15. Stereo decoders proved susceptible to noise on the supply line and filtering is needed to measure signal-to-noise ratios much above 60dB. Regulator should be mounted out of the transformer's magnetic hum field. 2000µF capacitor should have low internal resistance. ▼





▲ *Fig. 16. Circuit of a v.h.f. oscillator using a printed coil and providing a simply repeatable output level. Output voltage into 75 ohms is 55mV at 108MHz and 65mV at 87.5MHz. Temperature stability over 20 to 57°C at 96.4MHz is 4kHz / deg C. Deviation sensitivity at 104MHz relative to 88MHz is +5dB.*

The oscillator will run from either +12 or +15 volts so it can be run from the coder's supply or tapped from the receiver under test. The capacitor types used should be observed as they were chosen empirically to reduce the temperature drift. Wiring inside the box onto the p.c. board should use thin flexible wire with a slight slack left so that microphony is not transmitted from the input and output connectors onto the board.

The phase and amplitude mangling of the S signal which occurs in most receivers is so large that degradation is clearly visible on the demodulated composite signal even without any vertical magnification. Both low S amplitude and phase shift should be seen at 15kHz with S amplitude loss being predominant for 1kHz modulation. Oscilloscope synchronization will be helped by locking to the audio input to the coder or the deemphasized audio output from the receiver's active channel.

15kHz filter

This is just a convenient p.c. board which runs from 12 volts and will remove switching frequencies at decoder outputs without introducing significant distortion, so allowing distortion and signal-to-noise measurements. The resistor from pin6 to supply draws a small current to stop the crossover distortion which 741s otherwise generate with only a 6-0-6V supply. To make distortion measurements below about 0.15% two such filters are needed to completely remove ultrasonic components.

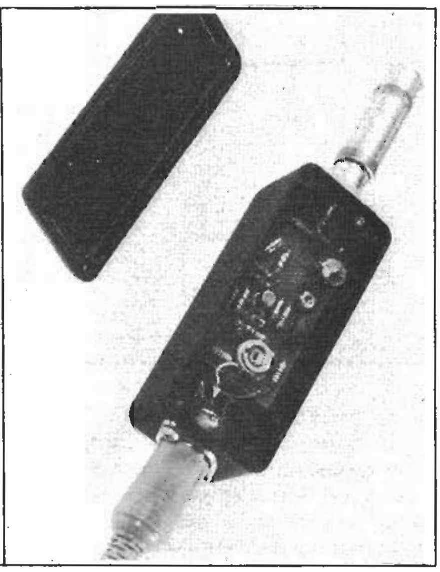
I think the coder design presented here has reached a cost/performance plateau. Many of its identifiable deficiencies can be attributed to the balanced modulator i.c., and £80 or so spent on a precision multiplier will provide some further improvements. The lack of inductors and single p.c. board make for a repeatable unit with stable performance.

The work described forms the basis of stereo coders for broadcast transmission, outside broadcast radio links and test units. □

Fig. 17. The v.h.f. oscillator shown just fitting into the smallest diecast box available (RS Components 509-923). Coaxial attenuators provide lower signal levels for receiver alignment. ▼

This series was written by . . .

Trevor Brook, who is keeping quiet for the time being about his latest idea, being a method of reducing noise in cassette tape machines he has decided to approach manufacturers with it first. But starting the electronic side of a new company to make film and tv equipment directly after leaving South London College (then Norwood Tech) must have convinced him that he could do the same sort of thing for himself, for he formed SurreyElectronics five years ago with a capital of £200. So we may see him making noise reduction modules as well as distribution and monitoring amplifiers, peak programme meters, and frequency shifters. His interests are not confined to audible frequencies. Acquiring a transmitting licence in 1966, he looked for good auroral openings by charting a tv sound channel from a transmitter 700km away, and heard a 20 watt repeater at Kilkeel over a 500km path "passing directly through Snowdon with unusual diffraction effects". With the aim of detecting sporadic-E backscatter and aurora he obtained a Home Office licence for an experimental pulse radar, but never quite overcame the problem of receiver blanking with a good noise figure.



▼ *Fig. 18. Circuit of a convenient filter for removing ultrasonic signals when making decoder measurements. Distortion at +11dB, 0.04%. Response −34dB at 19kHz, −45dB at 38kHz; ripple below 15kHz is less than 0.5dB. Crosstalk −80dB at 1kHz, −55dB at 15kHz. Noise −96 to −82dB over gain adjustment range of +4 to +14dB.*