F.M. Stereo Tuner

High-performance design for home construction

by L. Nelson-Jones, F.I.E.R.E.

In recent years there have been a number of developments in the components field, particularly in semiconductor devices, that have led to great improvements in the design possibilities for f.m. broadcast receivers. In particular these have been the advent of the dual-gate m.o.s.f.e.t., integrated-circuit i.f. amplifiers and demodulators, ceramic filters and improved variable-capacitance diodes. This two-part article describes an f.m. tuner design using these devices, discusses the advantages of the devices and gives constructional and alignment details. It does not attempt to be all-embracing and there will doubtless be some who disagree with the author's views of the current scene. It is hoped however that they do show some ways in which f.m. tuner design is currently evolving.

The work is the result of many months of design and measurement on five prototypes, so that results are not based on a one-off, and should be reproducible by readers who wish to copy the design. The receiver was designed to achieve in a relatively simple way a performance equal to the better examples of the commercial models available, but at a much lower price. (Total material cost comes out at about £11.) Comparison with the figures given in a recent *Wireless World* survey of commercial tuners (September 1970) suggest this aim has been achieved. The performance of the tuner under normal conditions of use has been excellent. One of the units is in use in Blandford Forum in Dorset—very much a fringe area—and gives noise-free reception from the Isle-of-Wight transmitters, including the new local station Radio Solent.

The design for the front-end of the f.m. receiver is shown in Fig. 1. Both r. f. amplifier and mixer stages use dual-gate f.e.ts with gate protection diodes. In the r.f. stage the upper gate is decoupled and acts as a screen between drain and gate 1, much as the g_2 electrode of a thermionic valve does. In the mixer stage this second gate is used as the injection point for the local oscillator voltage. There is not the same need for a screen between drain and gate 1 in a mixer stage as the drain load is not tuned to either the signal or oscillator frequencies. There is therefore little or no gain at signal frequencies to cause oscillation provided care is taken with the layout, particularly the length and placing of leads.

The magnitude of the local oscillator injection at gate 2 will affect the mixer gain and the spurious signal response characteristics of the mixer stage. This local oscillator voltage will be higher than in transistor tuners using bipolar devices by up to an order of magnitude and for the circuit conditions used a value of

Fig.1. Front-end of receiver using dual-gate f.e.ts.

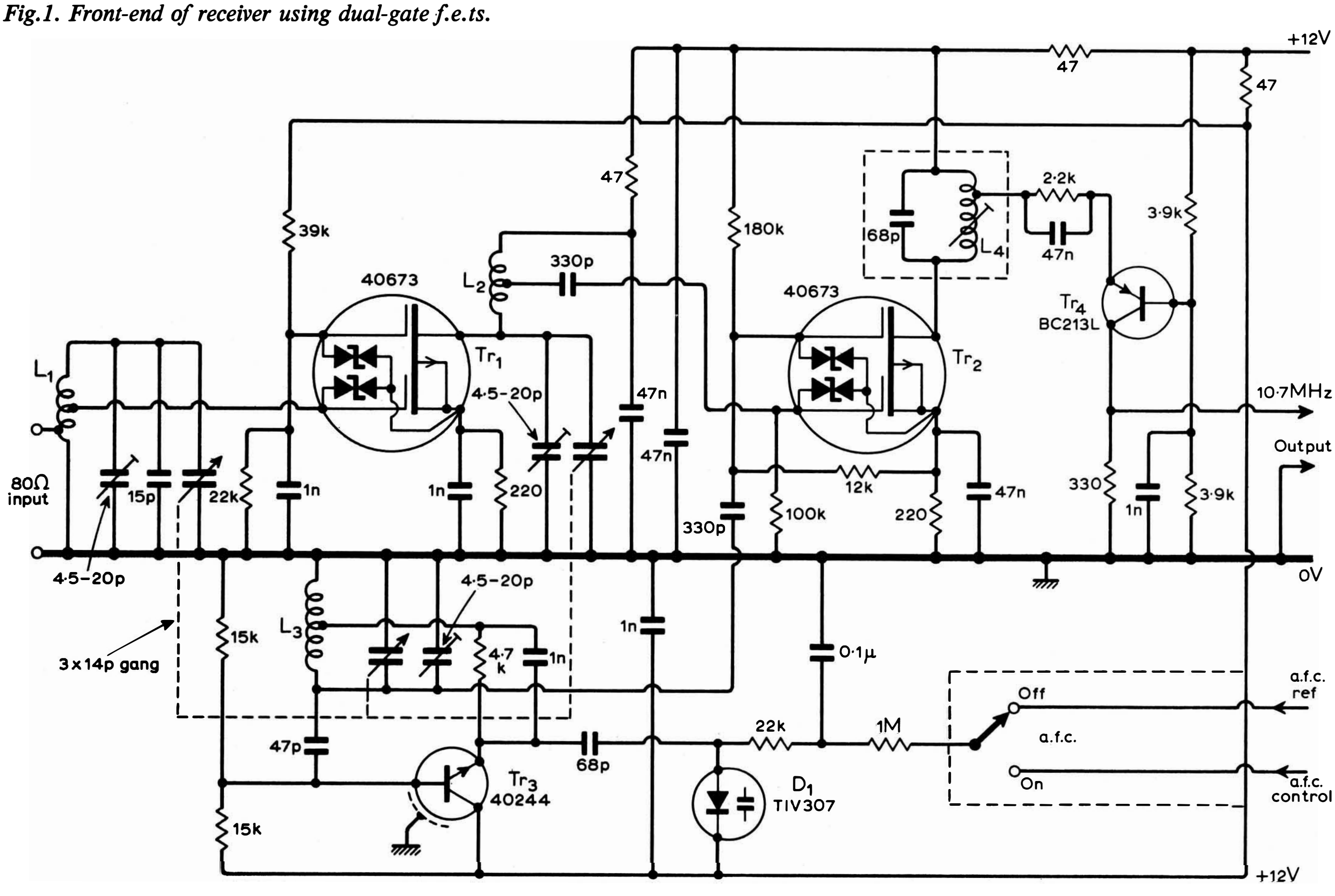
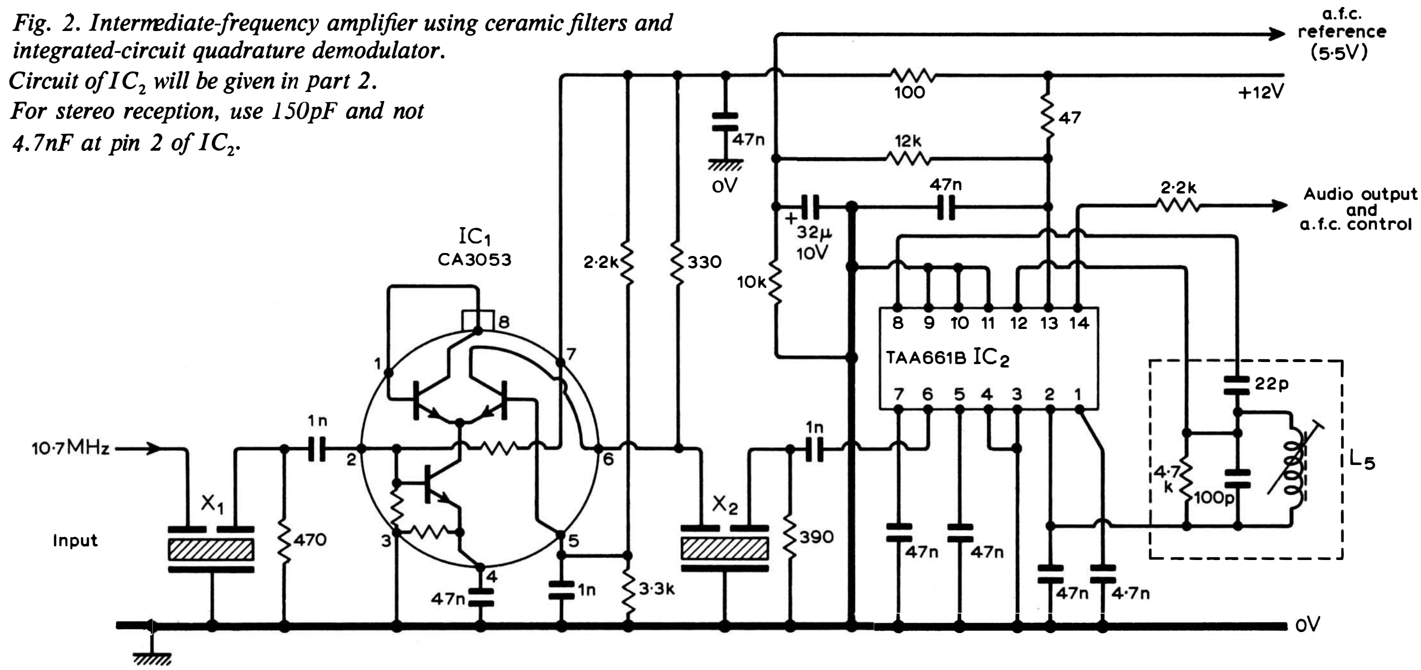


Fig. 2. *Intermediate-frequency amplifier using ceramic filters and integrated-circuit quadrature demodulator. Circuit of IC_2 will be given in part 2. For stereo reception, use 150pF and not 4.7nF at pin 2 of IC_2.*



500mV r.m.s. gives a reasonable mixer gain without too high a spurious signal response. In fact higher levels have been used without great trouble from spurious signals. Far greater trouble can be caused by lack of screening, leading to i.f. harmonics being picked up by the front end—especially with the high sensitivity of this tuner and, because of its small size, the close proximity of the front-end and the i.f. amplifier.

The oscillator is a conventional Hartley circuit with the ground point moved to give a grounded-collector design. There is no particular advantage to be gained in using an f.e.t. in this stage so that the cheaper bipolar device is preferred. Automatic frequency control is applied by the variable-capacitance diode D_1 coupled to the emitter of the oscillator stage. A resistor of 22 kΩ prevents the decoupling capacitor of 0.1μF from shorting out the oscillator voltage. The 0.1-μF capacitor together with the 1 megohm resistor form a low-pass filter to prevent audio voltages in the a.f.c. voltage from reducing the modulation of the carrier by audio frequency modulation of the local oscillator.

The a.f.c. can be switched out of operation by connecting the diode to a constant reference voltage from the i.f. amplifier. Diode D_1 is returned to the 12-volt supply line of the oscillator so that the a.f.c. control voltage changes the diode reverse voltage in the correct direction to reduce any oscillator drift. An increase in local oscillator frequency increases the intermediate frequency, which in turn leads to a rise in the output potential of the demodulator of the i.f. integrated circuit IC_2 (Fig. 2). As the diode is connected to the +12V supply, this increase reduces the reverse bias across D_1, increasing the diode capacitance and reducing the local oscillator frequency to correct its drift.

The mixer has a grounded-base stage feeding the 330-ohm resistor needed to correctly terminate the first filter unit X_1 (Fig. 2). This resistor also makes a convenient low-impedance output point from the front-end. A cheap p-n-p bipolar device is more than adequate for this position, because in a grounded-base configuration the requirements in respect of high frequency or noise performance are not stringent. The working Q of the tuned circuit is less than 20 so that tuning is not critical, and it is set to maximize gain in the usual way.

The supply for gate 2 of the r.f. stage is derived from the decoupled oscillator supply rather than from the top of L_2. This is brought about purely by layout convenience on the printed wiring board and, as gate 2 is additionally decoupled, has no effect on the performance.

I.F. amplifier

Two ceramic filter units X_1 & X_2 are separated by a buffer amplifier (IC_1) of moderate gain (about 20dB). The reason for this moderate gain is that it is desirable to place the filters as early as possible in the i.f. amplifier so that as successive stages limit with increasing signal level there is no change in bandwidth. This would be fully achieved if the whole of the i.f. gain were after the filter, and provided the mixer did not limit.

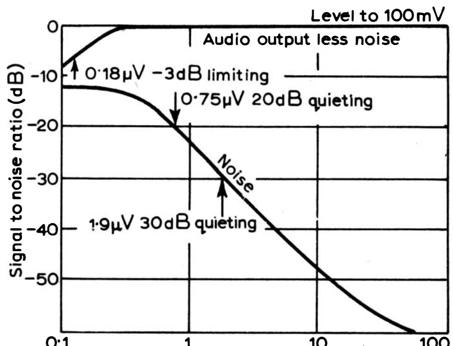


Fig. 3. Graph of low-level performance shows sensitivity of 0.75µV (for ±75kHz) for 20dB quieting. Above 50uV input, signal-to-noise ratio is better than 60dB.

In practice it is not possible to achieve this ideal, but the compromise of using only moderate gain before the second filter unit is a reasonable one and does not give rise to any undue increase in bandwidth over the normal range of signal levels. The performance obtained is a great improvement over normal bipolar i.f. amplifiers using discrete components with several double-tuned i.f. transformers. In such an amplifier the selectivity of the transformers is gradually lost starting at the output end as successive stages limit and the overall selectivity can leave a lot to be desired at high signal levels.

The first integrated circuit is a long-tailed pair circuit, used as a cascode amplifier by ignoring one of the top pair of transistors and driving into the long-tail transistor. The input impedance of this stage is suitable for the ceramic filter unit so far as resistance is concerned, but is above the maker's recommendations so far as input capacitance is concerned. For this reason the resistor terminating the filter X_1 is raised to 470 ohms, which compensates for the increased capacitance loading of the stage in restoring the top of the filter characteristic to reasonable flatness.

The load of the cascode stage is a 330-ohm resistor—to drive the filter X_2 from the correct source impedance. This low value results in the stage gain being low, especially when the loading effects of the filter are accounted for, so that although the slope of the cascode stage is around 100mA/V the overall gain from the output of X_1 to the input to X_2 is only a little over 20dB.

The input impedance of the IC_2, around 2 kΩ, is not very much greater than 330 ohms so that a terminating resistor of 390 ohms is used at the output of X_2. The value of the feed capacitor to the 'quadrature' tuned circuit L_5 is larger than the maker's recommended value of 18pF.

The value is not very critical in practice and up to 47pF has been used with little change in performance once the circuit had been retuned. It is likely that with large departures from the recommended value there will be an increase in distortion, but no appreciable effect is likely to stem from the increase to 22pF, a value more readily available than 18pF.

One possible reason for the apparent insensitivity to the value of this capacitor is that the quadrature drive voltage is derived from a resistive tap on the load resistor of the final stage of the i.f. amplifier of IC_2. The value of the impedance at this tap is fairly low, and as the coupling to the circuit is increased by increasing the value of the capacitor, the damping effect of this resistor increases, lowering the Q of the quadrature circuit, and compensating—at least to some extent—for the increased drive and tending to restore the correct phase relationship.

The audio output from pin 14 is taken via a 2.2-kΩ resistor to provide additional overload protection.

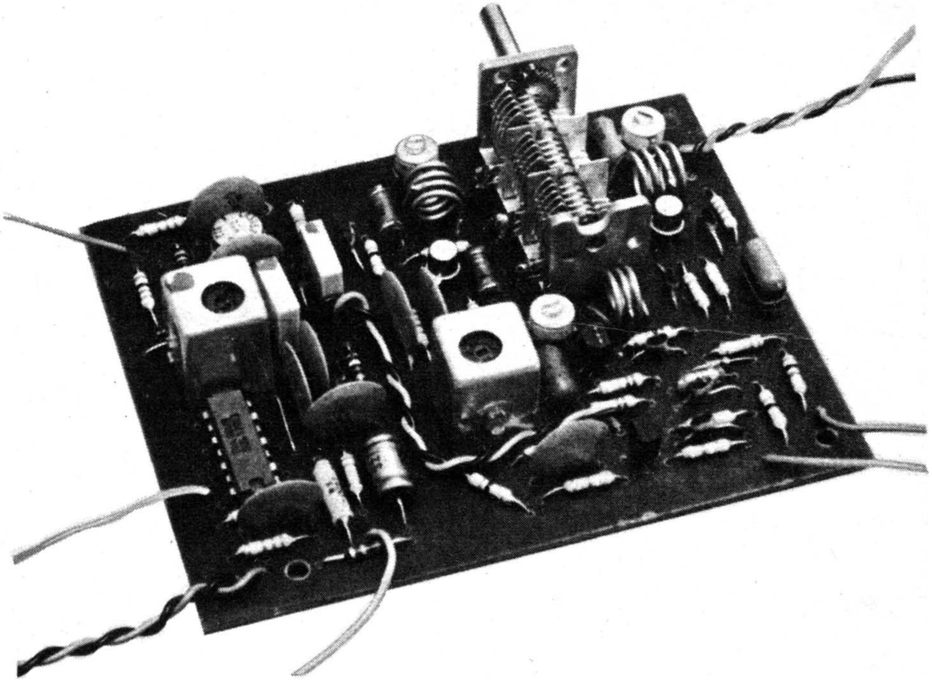
The reference used when a.f.c. is not required is derived from a potentiometer across the supply. The values chosen give a voltage close to that obtained at pin 14 of IC_2 (when there is either no signal or the signal is at centre frequency). The output level at pin 14 of IC_2 stays fairly close to around 46% of the supply voltage over the range 10-16 volts, and thus a simple potential divider is adequate for this reference voltage as this will also provide such a percentage of the supply voltage.

Because the output at pin 14 is a percentage of the supply voltage it is essential that the supply to the tuner be very well smoothed if hum and noise on the output is to be prevented. This is also important as the a.f.c. diode is returned to the 12-volt supply, and supply rail hum will therefore produce frequency modulation of the intermediate frequency whether or not a.f.c. is switched on.

The requirement for a ripple-free supply was one of the reasons for the choice of a 12-volt supply, so that adequate resistive smoothing could be used with a large reservoir capacitor. The drain of the tuner is almost constant and fairly well defined at a little under 35mA so that simple resistive dropping is satisfactory. An important point is that with such a network it is essential that the supply to the resistor be switched to disconnect the receiver, and not the supply to the receiver from the reservoir capacitor. In this latter case the capacitor would charge up to the full supply voltage with the receiver off, and on switching on the receiver would momentarily receive the full supply, which might well be enough to permanently damage the active devices of the receiver, especially IC_2. For supplies much above, say, 36 volts some form of simple stabilizer would be preferable.

Performance

The low-level performance of the tuner is shown in Fig. 3. At signals above 50μV



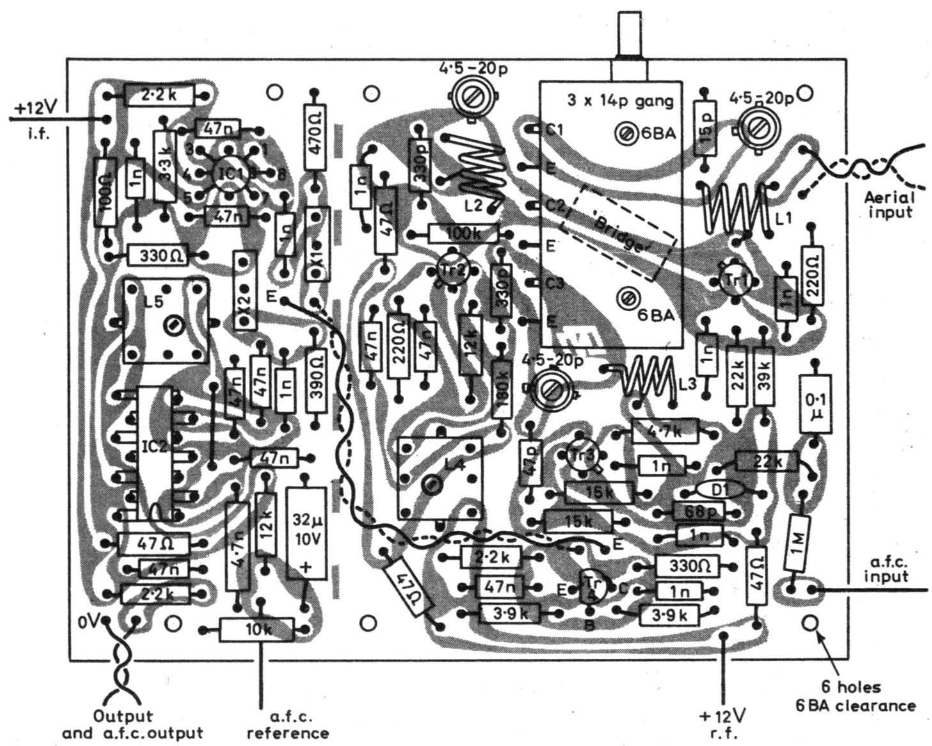


Fig. 4. Metal bridge—covering lead from Tr_1 drain to its tuned circuit—under the tuning capacitor is essential to maintain stability. Bridge, which can be tinplate, is detailed in Fig. 7. Complete tuner is screened by fitting into a die-cast box.

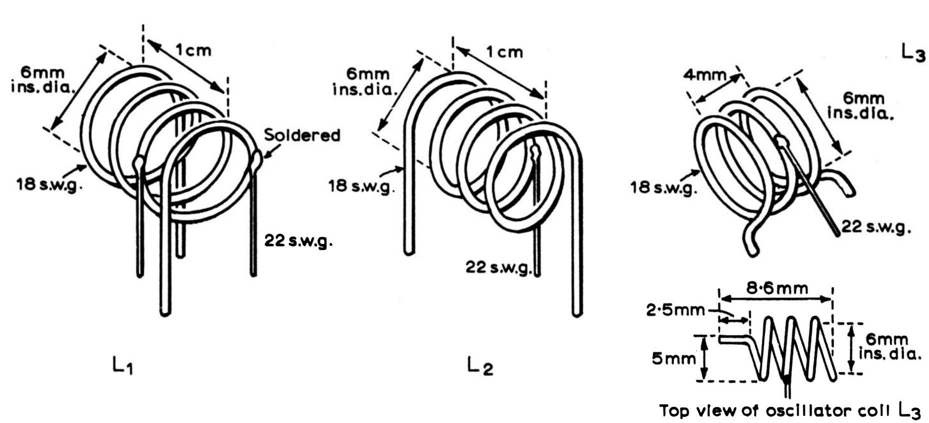


Fig. 5. Self-supporting r.f. coils, wound with slightly stretched wire, are shaped on a ¼-in dia. rod.

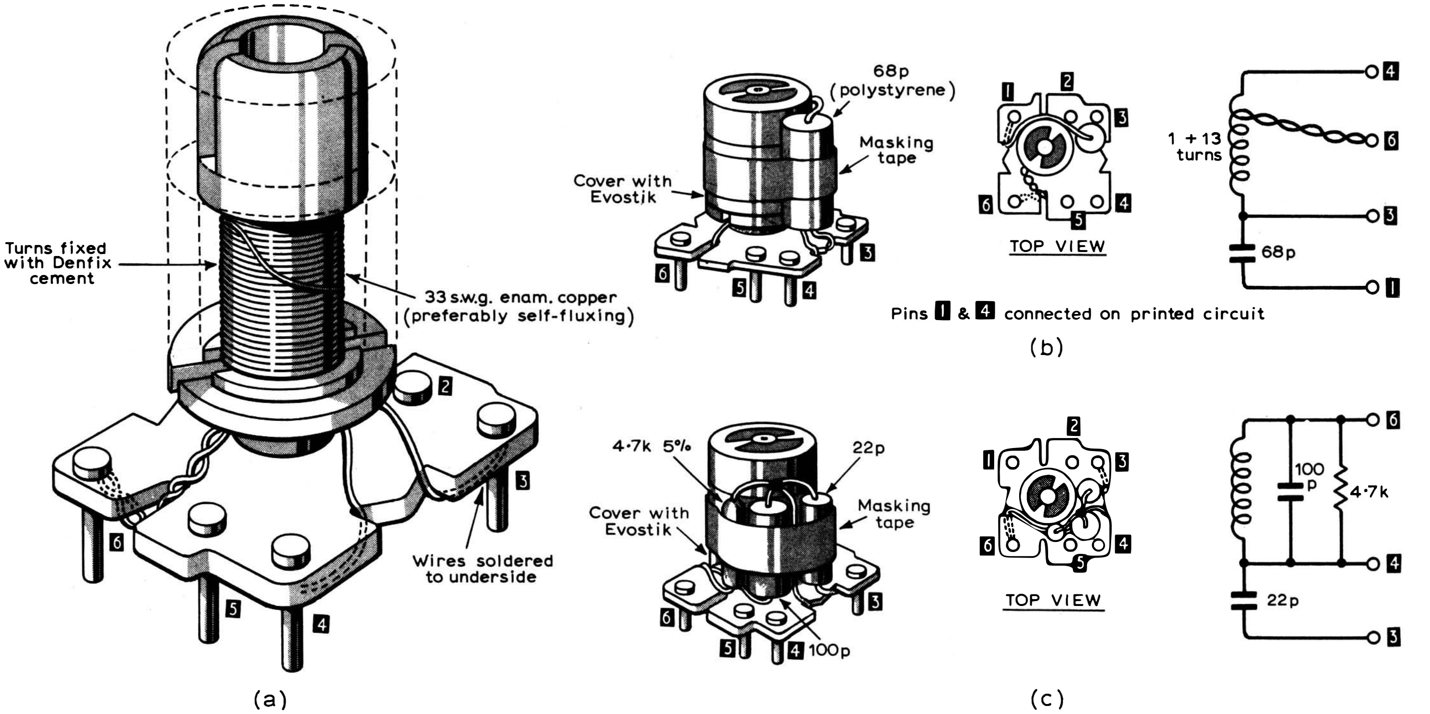


Fig. 6. Two screened i.f. coils are wound with 33-gauge enamelled wire and secured with a little Denfix cement. Capacitors are fixed with masking tape.

the signal-to-noise ratio is better than 60dB, and the 20 and 30dB quieting figures are 0.75 and 1.9μV respectively. Limiting (-3dB) of the demodulated audio signal (single-tone filtered from noise) occurs at only 0.18μV so that at all usable signal levels the i.f. section is limiting.

Unfortunately no signal generator was available which had an output with amplitude modulation-free from f.m., and it has therefore not been possible to check the a.m. rejection of the receiver. However, it is expected that the result will be close to the figure quoted for the i.f. integrated circuit (IC_2) at moderate signal levels, with an improvement when the first i.f. amplifier limits. The figure quoted for the TAA661B is 40dB at 10mV input, equivalent to around 10μV input at the aerial. Performance is summarized in the table on this page.

General layout

The tuner is constructed on a single-sided printed circuit board and is divided into two areas. The front-end and the i.f. amplifier are laid out separately, side by side, on a printed circuit board about 10 $\times$ 8 $\times$ 5cm overall, and in such a way that they may be separated if desired.

The complete tuner is enclosed in a screened box to cut down on spurious responses due to radiation from the i.f. amplifier, and to reduce local oscillator radiation to a minimum. This screening is especially necessary with this design because of its very high sensitivity.

The dial drive system suggested gives a scale length of 13.7cm with a reasonably linear frequency change over the centre 80% of this scale. A cord drive system is used which has the advantage of retaining the pointer at both ends, thus eliminating the problem of sliding friction at one end of the pointer, and giving a much smoother drive.

The overall layout of the components is shown in Fig. 4 as seen from the components side of the board, and also in the oblique view.

In construction keep leads short and if possible test all components on a bridge before fitting them on the board, as this can save ruining a good p.c. board should the tuner not work first time. It cannot be emphasized too strongly that such component checks can save much wasted money and temper. It is also vital to check that the components are correctly located, the diode connected with the right polarity, and that there are no breaks or shorts on the 'track' of the p.c. board. This latter point is of importance on such a small board with roughly 200 component holes, as tracks are necessarily fine and gaps small. A watchmaker's eyeglass has been the constant companion of the author during the construction and design of this tuner.

Coil construction

The r.f. coils are all made from 18-gauge tinned copper wire and are self-supporting. Taps are made by soldering leads of 22-gauge tinned copper direct to the turns of the coils—Fig. 5. The coils were made by winding the wire on a $\frac{1}{4}$-in rod such as a drill shank. The wire should be straightened by placing one end of a length in the vice and pulling the other end until the wire stretches very slightly, when the wire will have lost all kinks.

The coil wires should be a firm push fit in the board, if undue strain on the copper foil of the board is to be avoided when adjusting the coils. At all cost avoid the wires being loose in the board before soldering and if necessary apply the minimum of Araldite epoxy adhesive around the 18-gauge coil wires on the components side of the board after soldering in position. The joints should then be quickly reheated with the soldering iron to cause the Araldite to run into the holes in the board, thus securing the coils rigidly. After using such an adhesive the board must be left in a slightly warm place, e.g. an airing cupboard, for 24 hours to ensure that the adhesive has set hard.

The coils should be mounted on the board in the positions shown in Fig. 4, and with the turns of the coils nearest the board surface 2.5mm clear of the board. In the case of the oscillator coil this must also be 3.5mm clear of the rear face of the tuning capacitor. It is best to adjust the coils before soldering them into the circuit board to minimize subsequent adjustment, and the overall coil lengths given (over the outside of the end turns) are close to the

Performance

Sensitivity	
–3dB limiting	0.18μV
20dB quieting	0.75μV
30dB quieting	1.9μV
Spurious response	
image rejection	–70dB
i.f. rejection	–85dB
other unwanted signals	–94dB
Audio output	0.5V r.m.s. for $\pm$75kHz
*Capture ratio**	2dB approx.

* Difficult to measure. There does not appear to be much dependence on signal level provided the signals are reasonably above noise level—a result to be expected from the very low level at which limiting starts. Figures varying from 1 to 4dB were measured at various signal strengths on repeated measurements. In general a signal 10dB below produced no noticeable effect on the demodulated output.

final adjusted lengths required for correct tracking.

An alternative method of mounting is to open up the main coil mounting holes in the board and insert 'eyelets' which are big enough to allow the 18-gauge coil leads to pass through them. The eyelets are riveted into the board so that the strain is removed from the copper track. Overall connection is obtained by soldering the track to the eyelet and to the coil leads. All the main pads for the ends of the coils are large enough to allow this to be done.

The two screened i.f. coils are both constructed on Neosid coils type NS/E3, and both are wound with 33-gauge enamelled wire, preferably of the self-fluxing variety. The two coils are wound as shown in Fig. 6 and the turns secured in place for stability with a minute quantity of Denco Denfix polystyrene cement. (Do not use modelmaker's polystyrene cement as some varieties can have high loss factors.) It is essential to use the least possible quantity as the bobbin of the coils is made from polystyrene loaded with iron dust and is very easily dissolved by this cement. When the cement has dried, place the ferrite sleeve over the coils, ensuring that the leads are well pressed down in the slot at the base of the bobbin so that this ferrite sleeve does not scratch the wires.

Next push on the polythene retaining disc. Secure the ferrite sleeve to the coil former with a smear of Evostik latex-resin contact adhesive around the join between the sleeve and the coil former near the base. When dry connect the capacitors (and the resistor in the case of L_5). Tape these components to the former as shown to hold them clear of the coil can. Make sure all leads are well clear of the can when this is slipped over the coil. Next fix the core into the coil. The ferrite core cuts its own thread into the polythene top retainer. Set the top of the core level with the top of the polythene retainer–Fig. 6b. This should be close to the final adjustment position. In the case of L_4 the capacitor is connected to the coil on one side via the printed circuit, which connects pins 1 and 4, thus placing the 68-pF capacitor across the coil.

Fitting components to the board

Due to the small size of the board and components it is absolutely essential to use a soldering iron with a small tip which is adequately hot and clean. Lead lengths must be kept short and the components close to the board. This is especially important with the ceramic disc decoupling capacitors (1 and 47nF types). The transistors should be pushed down onto the board until the body is 2.5mm above the board; pushing the transistors closer than this will strain the leads unnecessarily. This rule applies also to IC_1. The second i.c. should be placed down on the board until the shoulders on each lead contact the board; the body of the i.c. will then be just clear of the board.

There will be some difficulty in locating the polarity of the diode due to the small size of this device. If in doubt check it with

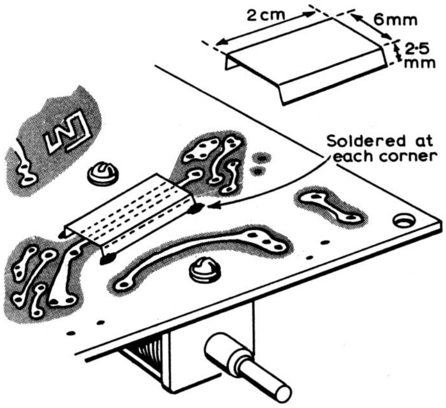


Fig. 7. Metal bridge—covering lead from Tr_1 drain to its tuned circuit—under the tuning capacitor is essential to maintain stability.

an ohmmeter on the ohms × 1 range, when the cathode will be the one connected to the positive lead of the multimeter when the meter shows conduction. (On a multimeter the polarity of the leads on the resistance ranges is the opposite to that shown on the meter panel so that the red positive lead is negative, and when connected to the cathode results in the diode being forward biased.)

There are two wire links on the board, one on either side of L_5 in the i.f. section. These links may be either 22-gauge tinned copper or 1-024 p.v.c. covered wire. Connections to the 3-gang capacitor are by similar wires. The capacitor is secured to the board by two 6BA screws not longer than 4.5mm of thread.

The link from the r.f. to i.f. sections is by a twisted pair of 1-024 p.v.c. insulated wire as shown in Fig. 4 and in the oblique view. Take care to see that this is correctly connected, i.e. the live lead of the pair connects between the collector of Tr_4 and the input to X_1.

The two screened coils L_4 and L_5 are soldered into circuit after being pushed well down on the circuit board so that when the can is placed over the coil it just rests on the polythene retainer at the top of the coil, while also just contacting the p.c. board. The can is put on the coil after soldering the coil into the board, and the two can tags are then also soldered to the board.

There is one component not shown on the circuit because it is not a circuit component. This is a 'bridge' across the lead from the drain of Tr_1 to its tuned circuit under the tuning capacitor. This bridge continues the earth plane as well as screening the lead. It is essential to use this bridge to maintain stability in the r.f. amplifier. The bridge is necessary because of the layout limitations set by the capacitor having its connections on only one side of the body, and the need to keep the coils well spaced to obtain good stability and keep oscillator radiation low.

The dimensions and location of this bridge are shown in Fig. 7. The bridge may be made of any metal that will not corrode but will solder. Tinplate was used in the original units. Take care not to short the wire to the centre-section stator of the variable capacitor at the end of the bridge nearest to L_2.

If the ceramic filters need removing from the board, take care not to apply pressure when applying heat to the connections, otherwise the component will be damaged. Remove solder first with a desoldering tool or with copper braid.

In the photograph the alternative type of oscillator transistor is shown. If this type is used then the fourth connection on the transistor won't exist—used to earth the can on the TO-72 type specified (40244). Only the three connections nearest the 3-gang capacitor will then be used. In addition, an extra lead is shown in the photograph adjacent to the integrated circuit IC_2 that is not shown in Fig. 4. This lead connects to pin 14 of this i.c. to control the a.f.c. However, as the output lead (via the 2.2-kohm resistor) will

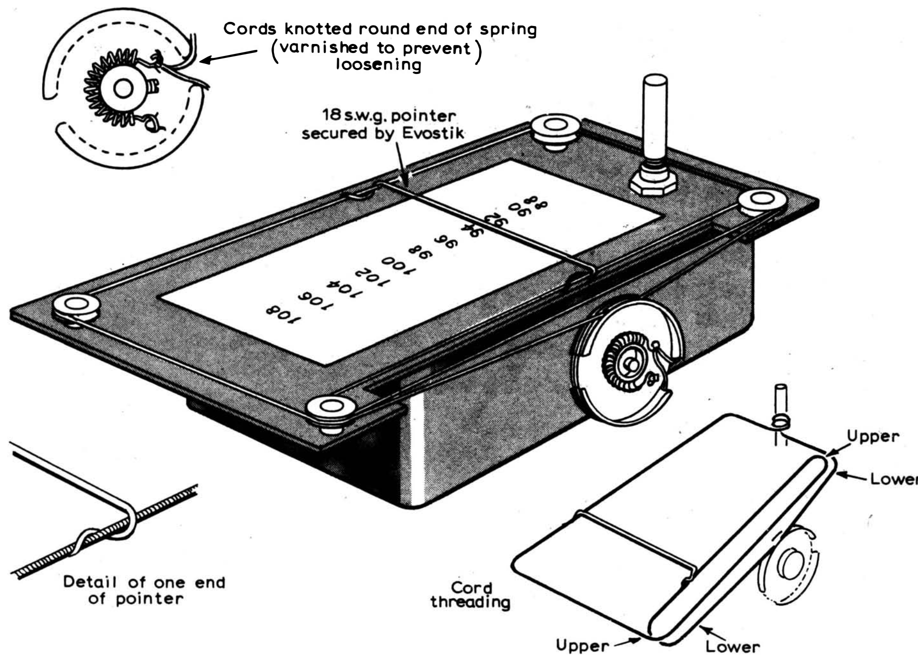


Fig. 8. Suggested cord system eliminates pointer friction and can be mounted in any plane.

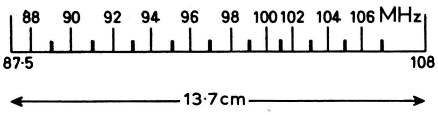


Fig. 9. Typical scale graduations for the band 87.5 to 108 MHz.

normally have a d.c. blocking capacitor in series, it will be at the same potential as pin 14, and is therefore suitable for the same purpose.

Screening of the tuner

It is essential to screen the tuner to avoid instability in the i.f. section due to pick-up, particularly from the output lead. Although the de-emphasis capacitor removes most of the 10.7-MHz signal and its harmonics, the output lead can still have sufficient of these signals present to cause spurious whistles when tuning if this lead is anywhere near the r.f. section. A great improvement results from connecting a capacitor of 470pF from the output to ground which, with the 2.2k Ω series resistor, removes these high-frequency signals sufficiently to make the position of this lead less of a problem. (In stereo applications the capacitor used will probably need reducing to 100-200pF to avoid attenuating the stereo switching waveform unduly, and if a long screened lead is used for the output the capacitance of this lead should be enough by itself.)

Prototypes were fitted into an ITT die-cast case with internal dimensions 12 × 9.5 × 2.5cm and is a tight fit with regard to height. Connections to the tuner should be made through insulated feed-through terminals close to the board so that all leads are as short as possible inside the screened box. A slot will have to be cut in the side of the box to allow the tuning capacitor shaft to pass through. The author found it easiest to fit the tuner board to the lid of the box, and to use the box as a cover, with holes drilled in line with the trimmer capacitors and L_4 and L_5 to enable final alignment with the box closed.

The board is mounted by four 6BA screws as shown in Fig. 4 and must be spaced about 5-8mm from the surface of the screened box, to prevent the track-side of the board from shorting on the box. Extra nuts or spacers may be used to achieve this spacing.

Dial drive system

Fig. 8 shows the layout of the suggested cord drive system which eliminates the problem of pointer friction, and is suitable for mounting in any plane. The parts required are made by the manufacturers of the 3-gang capacitor with the exception of the pointer, made of 18-gauge tinned copper, or similar, and the cord. Typical scale graduations for the 87.5 to 108-MHz band are shown to scale in Fig. 9.

To be concluded with a discussion of devices used and alignment methods.

Parts list

Set of parts is available from Integrex Ltd, P.O. Box 45, Derby DE1 1TW.

Inductors
See illustrations and text for winding details L_4 & L_5 inductor assemblies are Neosid NS/E3

Dial components (all Jackson)
$\frac{1}{2}$-in brass pulleys, type 4879 (4 off)
$\frac{1}{2}$-in brass pivots, type 4539 (4 off)
brass spacers, type 4880 (4 off)
type 'G' drive spindle, type 5080
drive drum 2.5cm dia. 4-mm bore

Variable capacitors
3 × 14pF tuning capacitor, part no.5560/3/14 (Jackson)
4.5 to 20pF trimmer (3 off) (Piher make from Henry's Radio or Rosenthal type STSE-7, N750 from Radio Resistor Co. or type 7S-Triko 02 from Steatite Insulations)

Fixed capacitors
1nF disc ceramics, 50V, 1-cm mounting centres (10 off)
0.1μF, 16V (Mullard type C280)
32μF, 10V (Mullard type C426)
22, 68 (2 off) & 100pF, 160V, polystyrene
15, 47 & (2 off) 330pF ceramic tubular or disc, or polystyrene mounting centres 1cm
4.7nF miniature tubular polystyrene 1.65-cm mounting centres. Use 150pF for stereo reception

Resistors
Miniature carbon film type, $\frac{1}{8}$ watt $\pm$5% tolerance (Mullard)

Active devices
40673 (RCA 2 off). In mixer stage, lack of gate protection diodes may be acceptable, in which case types 40604 or 3N141 can be used. If the risk of not using diodes is acceptable for r.f. stage, types 40603 or 3N140 can be used. N.B.: retain protective spring until power is applied
40244 (RCA), alternatively TI409 (TO-92) —now avaialable as TIS64 (TO-18), Texas
BC213L (Texas) or BCY70
CA3053, 3028A or 3028B (RCA)
TAA661B (SGS)
TIV307 (Texas)

Also needed are
printed circuit board (available drilled, solder coated and with component locations)
ceramic filters type FM-4 (Vernitron). Order as pair with same colour coding (orange-10.625 MHz, yellow-10.6625 MHz, green-10.700MHz, blue-10.7375MHz, violet-10.775MHz)
trimming tool for L_4 and L_5 cores
nylon cord
die-cast box
Denfix cement
Denfix cement (from Home Radio)
Evostik latex-resin impact adhesive

Baird's Video Disc turns again

The video disc is not a new idea, as J. C. G. Gilbert pointed out in his article on the Teldec system last year.* John Logie Baird recorded video signals from his 30-line television camera on a 78 r.p.m. wax disc as far back as July 1928. Copies were made and sold to the public by Selfridge's store, London, in the early 1930s, the idea being that they should be played from an electric gramophone into the Nipkow-disc Baird Televisor of the time.

Recently *Wireless World* was able to examine one of these discs, and see the kind of pictures it produced on a Televisor, at a demonstration put on by the I.T.A. at the television museum in its London headquarters. The disc, which looks like a 10-inch black gramophone record with a red label, was acquired from Mr. G. Diment of Herne Hill, London, who bought it from Selfridge's in 1935 at a price of 7 shillings. On the label are the words 'Recorded Television Record No. 1, Speed 78, Scanning Speed 750, Lines 30, For Private Use Only'. Because the disc is very fragile a magnetic tape recording had been made of its video signals, and it was this taped copy which was played into the Televisor at the demonstration.

The Televisor was a 40-year old model, one of two acquired by the museum and restored to working order by I.T.A. engineers at the Fremont Point transmitter, Jersey. The authenticity of the results was assured by H. J. Barton-Chapple, who worked with Baird, and by P. J. Packman, who built one of the two Televisors in 1928 at Plessey.

All that can be said of the pictures seen is that they were a sequence of patterns, in the characteristic orange light of the Televisor's neon tube. What the patterns depicted was anybody's guess, though we were told they were caricatures of human faces. Certainly, the first video disc cannot be regarded as anything more than a technical curiosity. It is nonetheless a further tribute to Baird's ingenuity and enterprise in the face of the public's indifference at that time.

*"The Video Disc", *Wireless World*, August 1970, p.377.

Correction

Peter Blomley, author of the articles '**New approach to class B amplifier design**' (February and March issues), tells us Tr_5 in Fig.1 of the second article should be type 2N3904 and not 2N3905, and that in Fig. 5 the ordinate should be labelled 0.00075%/cm, and not 0.0012%/cm.

F.M. Stereo Tuner

2—Further details of high-performance design for home construction

by L. Nelson-Jones, F.I.E.R.E.

This sensitive f.m. tuner design, described in last month's issue, has a performance equal to the better examples of commercial tuners, but at a much lower cost. Full constructional details were given in Part 1 and this article discusses in detail some of the devices used—especially the dual-gate m.o.s.f.e.t., integrated circuit demodulator and ceramic i.f. filters—and concludes with alignment instructions.

The dual-gate m.o.s.f.e.t. is not to be confused with the type of junction f.e.t. which has two gate connections, usually one to the gate and the other to the substrate, as this has gates effectively in parallel. The dual-gate m.o.s.f.e.t. has gates effectively in series so that it can be likened to the multi-grid valve or a cascode stage and like these devices has the advantage of very low feedback capacitance from output to input. It has also the same advantages as single-gate m.o.s.f.e.t.s namely, good signal handling, low noise, and high input impedance. Fig. 9 shows the likeness of the dual-gate m.o.s.f.e.t. to a cascode stage, and its construction. The drain current of a dual-gate m.o.s.f.e.t. is a function of both gate potentials, and this enables gate 2 to be used for gain control in the case of r.f. amplifiers, or for injection of local oscillator voltages in the case of mixer stages. Type 40673 is very similar to the 3N140 but in addition has full protection of both gates by pairs of zener diodes between each gate and the source (and substrate) electrodes. These diodes are clearly of minute proportions—they add only a fraction of a picofarad to the gate capacitances. The breakdown of these diodes is around ± 10 volts, so that normal signal levels do not cause conduction. But the diodes will conduct long before the gate breakdown voltages are reached, and, provided the resultant currents are adequately limited by the circuit values, no harm will result to the gates.

Apart from the obvious advantage of two controlling gates, the great advantage of the second gate is that it acts as a 'guard ring' between the drain, and gate 1. The result of this guard ring action is a typical drain-to-gate 1 capacitance of 0.02pF (with a maximum for the 40673 and 3N140 types of 0.03PF). This low value of feedback capacitance enables such a device to give up to 28dB of power gain at 100MHz, without need for neutralization, but in practice a gain of 20dB is a more realistic figure for an r.f. amplifier at this frequency. This ensures a high margin of stability, which together with the superior signal handling qualities of the m.o.s.f.e.t. make this a very easy device to use for r.f. amplification in an f.m. tuner.

Integrated circuit i.f. amplifiers

Integrated-circuit i.f. amplifiers have been available for some time now in various forms, from the simple differential pair and the cascode stage, up to relatively complex circuits such as that used in the receiver described (TAA661B). There are now a number of these more complex circuits available, nearly all of which use a product detector for demodulation. Examples of these are the Sprague ULN-2111, Plessey SL432A, and the SGS TAA661B. Fig. 10 shows the circuit of the TAA661B, together with the basic external connections.

Gain is provided by three stages, each of which is a non-saturating differential amplifier followed by an emitter follower.

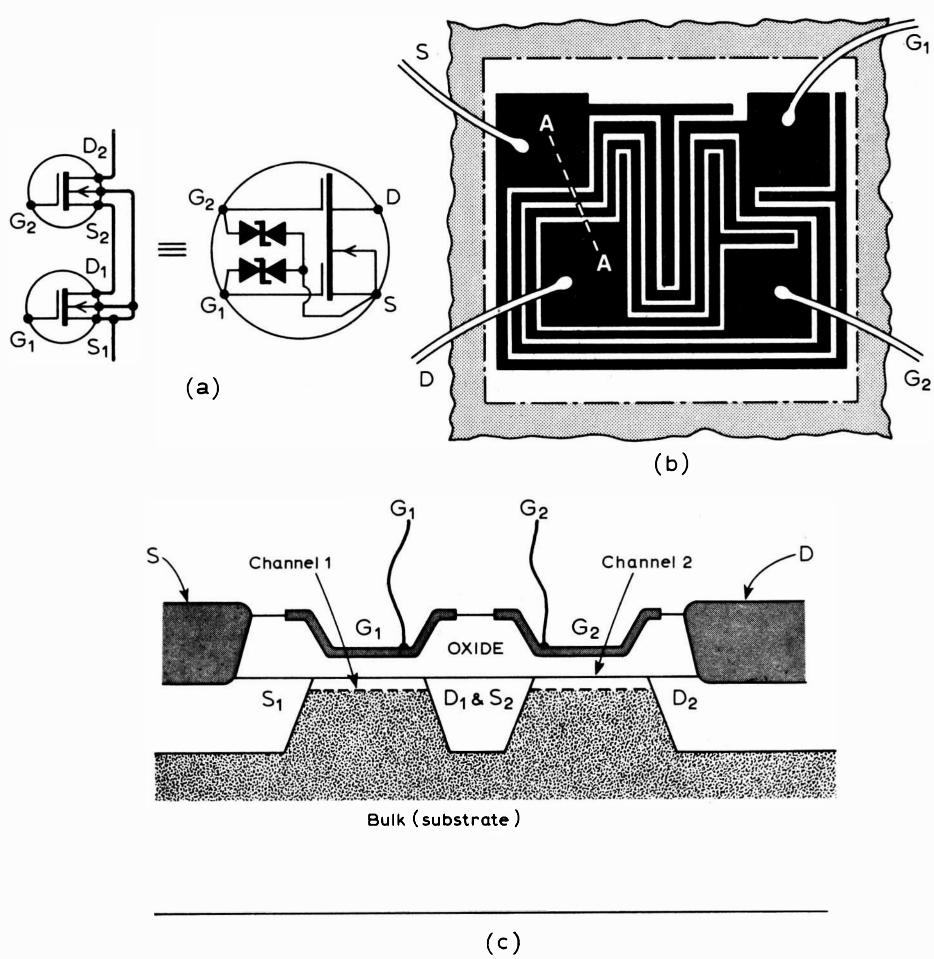


Fig. 9. Symbolic representation of dual-gate m.o.s.f.e.t. (a) showing similarity to cascode stage. Plan view (b) shows complete separation of gate 1 from drain-by-gate 2; (c) shows section across A-A. Bi-directional zener diodes conduct at around $\pm 10V$ preventing gate breakdown (type 40673 only).

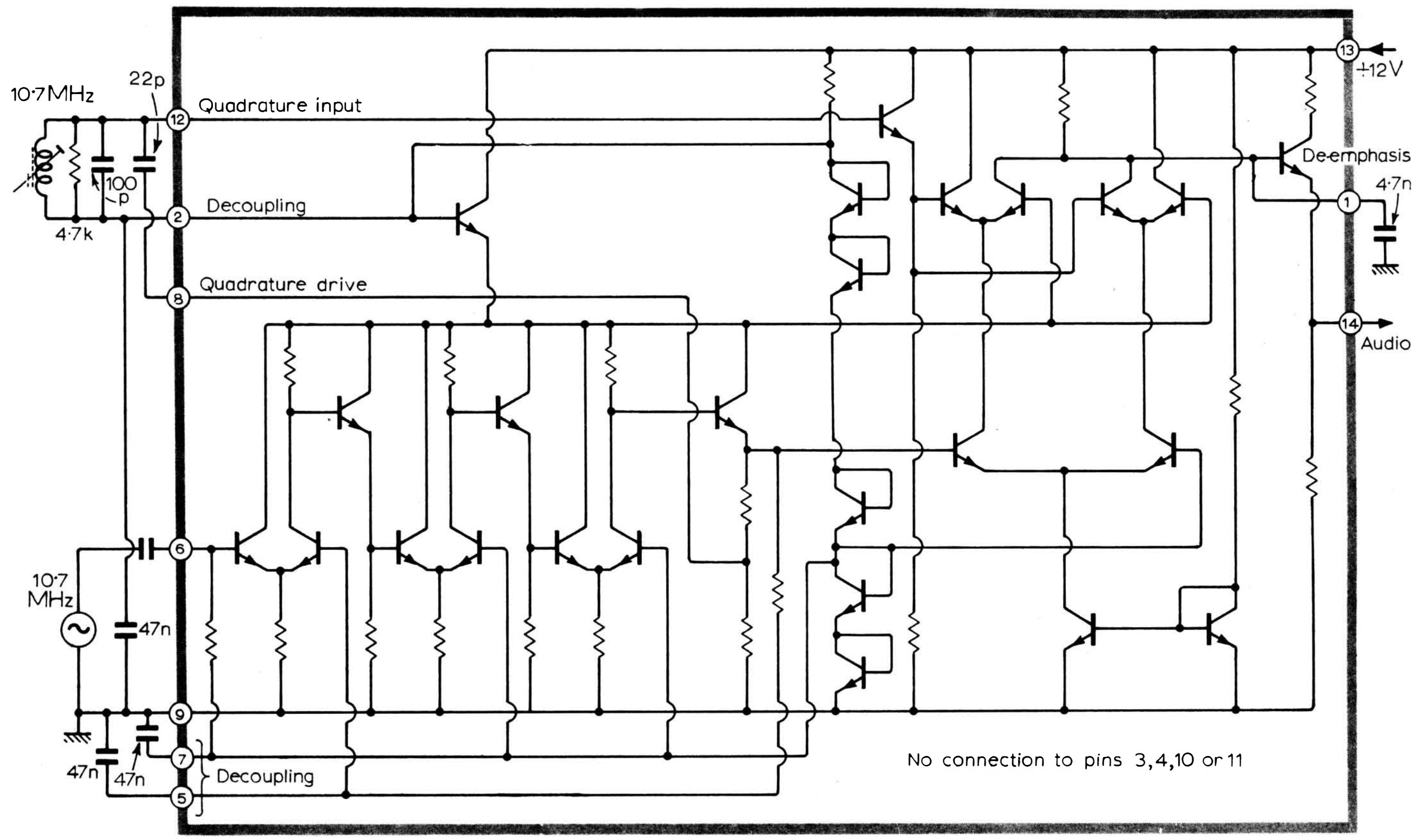


Fig. 10. Demodulation and i.f. amplification are performed in this single-chip integrated circuit (TAA661B). Phase-sensitive detector consists of 'tree' of differential pairs with constant-current tail (to right of bias chain), fed with a phase reference provided by tuned circuit and with signal to lower pair.

Overall d.c. feedback is applied so that the output level at the third emitter follower is kept equal to that of the base of the input transistor. This voltage is set at approximately 1.4 volts by the bias chain of five diodes which has two outputs, equal to two and five 'diode voltage drops'. The higher of these voltages is used to control the main supply line of the amplifier stages via an emitter follower. This supply line is therefore at approximately 2.8 volts—five 'diode drops' less the base-diode drop of the emitter follower.

The detector consists of a 'tree' of differential pairs with a constant-current source in the common 'tail' connection. This constant-current source is a 'current mirror' circuit where the constant current is equal to the current feeding the second transistor, strapped as a diode. The current mirror principle is based on the fact that two equal transistors with equal base-emitter voltages will also have equal collector currents. This principle may be extended so that two transistors of an integrated circuit having different areas (but otherwise similar) will have collector currents equal to their areas when used in such a circuit.

The detector acts as a phase-sensitive full-wave rectifier, with a phase reference provided by a tuned circuit driven from a tap on the load of the final emitter follower of the amplifier. The lower two transistors of the tree are driven by the signal from

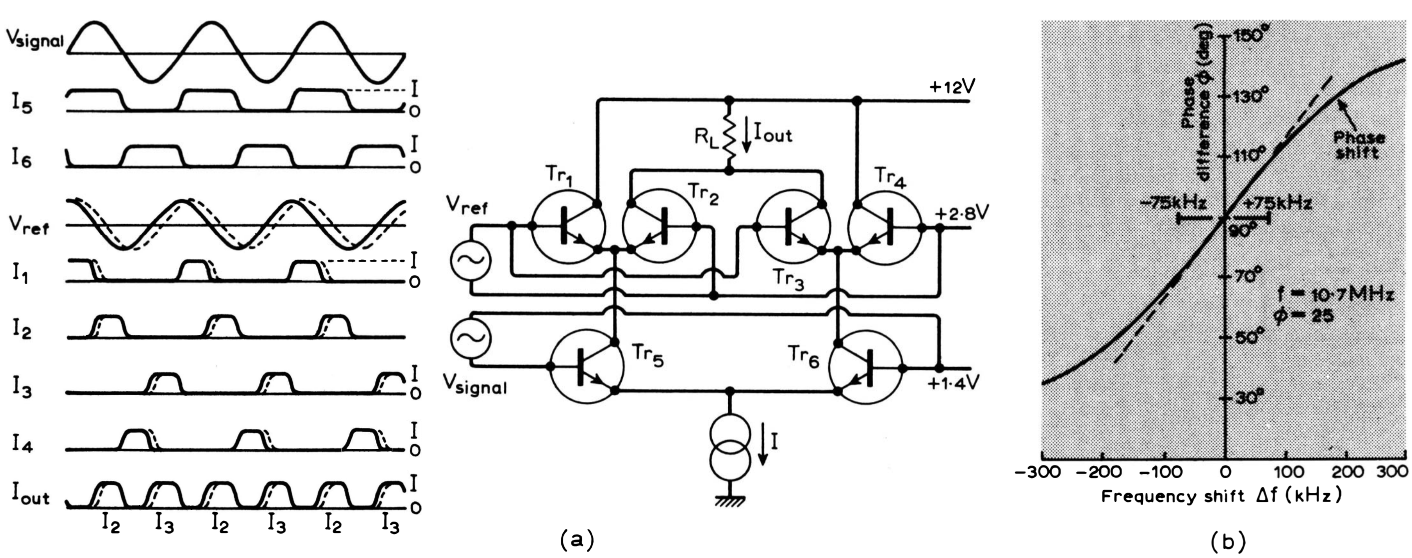


Fig. 11. At resonant frequency of tuned circuit—see Fig. 10—signal applied to lower differential pair is in quadrature with reference from tuned circuit fed to upper pairs, and current divided equally between each of the upper pairs (a). When signal frequency deviates, phase difference between the two signals increases or decreases (b), changing proportion of current through each half of both upper pairs.

the amplifier, and as the two bases of the pair are at equal d.c. potential—due to the overall 100% d.c. feedback over the amplifier—the collector currents of these two transistors become square waves at the carrier frequency, at all signal levels above the limiting threshold of the amplifier chain. The two upper pairs of transistors are fed by the reference voltage from the tuned circuit and like the lower pair the bases of both pairs are at equal d.c. potential.

One base of each pair is connected to the supply line of the amplifier, while the other is fed by an emitter follower, biased via the tuned circuit from the same potential as the base of the emitter follower controlling the supply line of the amplifier. The voltage across the tuned circuit is approximately 300mV peak-to-peak with full limiting so that these upper pairs of transistors are also fully switched at the reference frequency. At the resonant frequency of the reference circuit, the signal voltage applied to the lower pair of transistors is in quadrature with the signal from the tuned circuit to the upper pairs of transistors, due to the loose coupling of the tuned circuit via the 22-pF capacitor. Thus at resonance the current square wave through each half of the lower pair of transistors will divide equally between each of the upper pairs, because the two signals are in quadrature, and the transition of the reference waveform takes place midway through each half cycle of the current square wave supplied by the lower pair. Action is shown in Fig. 11a.

As the frequency departs from the centre frequency of the tuned circuit, the phase difference between the two signals decreases or increases, depending on the direction of the frequency shift, so that the proportion of the current passing through each half of each upper pair changes, Fig. 11b. The collectors of the upper pairs are connected so that the pair which have an increase in current for an increase in frequency are connected together, as are those having a decrease. One pair of collectors is connected to a load resistor, and the other pair direct to the 12-volt supply. The load resistor drops approximately 6.0 volts at the centre frequency so that the output level is typically +5.5 volts, at the emitter of the emitter-follower output stage.

De-emphasis is arranged by a capacitor connected to the base of the emitter follower. Alternatively a separate de-emphasis network can be connected to the output in the usual way, with a much smaller value of capacitor connected to pin 1. A capacitor connected to this pin is still essential to preserve overall stability by by-passing the r.f. voltages present at this point. A similar reduction is necessary if the output is applied to a stereo decoder. A value of 150pF is suitable in either case.

Ceramic i.f. resonators

There are a number of ceramic resonators on the market and they take different physical forms. Some are similar to the type of filter common in communications receivers where high degrees of selectivity are required, and these have tuned circuits at input and output with one or more resonators between. Others use only ceramic resonators, with perhaps coupling capacitors, and have a family resemblance to the type of crystal filter used in v.h.f. communications receivers for high degrees of i.f. selectivity at 10.7MHz. The type of filter used in the tuner consists only of a single ceramic resonator which by the layout of its electrodes performs the function of a multi-section filter with a bandpass characteristic.

Such filters are now also being made in quartz for v.h.f. communications receivers and can equal the performance of much more complex multi-element filters, despite their relative simplicity. This excellence of performance is true also of the ceramic type, where the device used has a performance slightly better than a multi-element device of otherwise similar characteristics, both in respect of selectivity and passband loss. Due to its greater simplicity it is also much cheaper, and smaller.

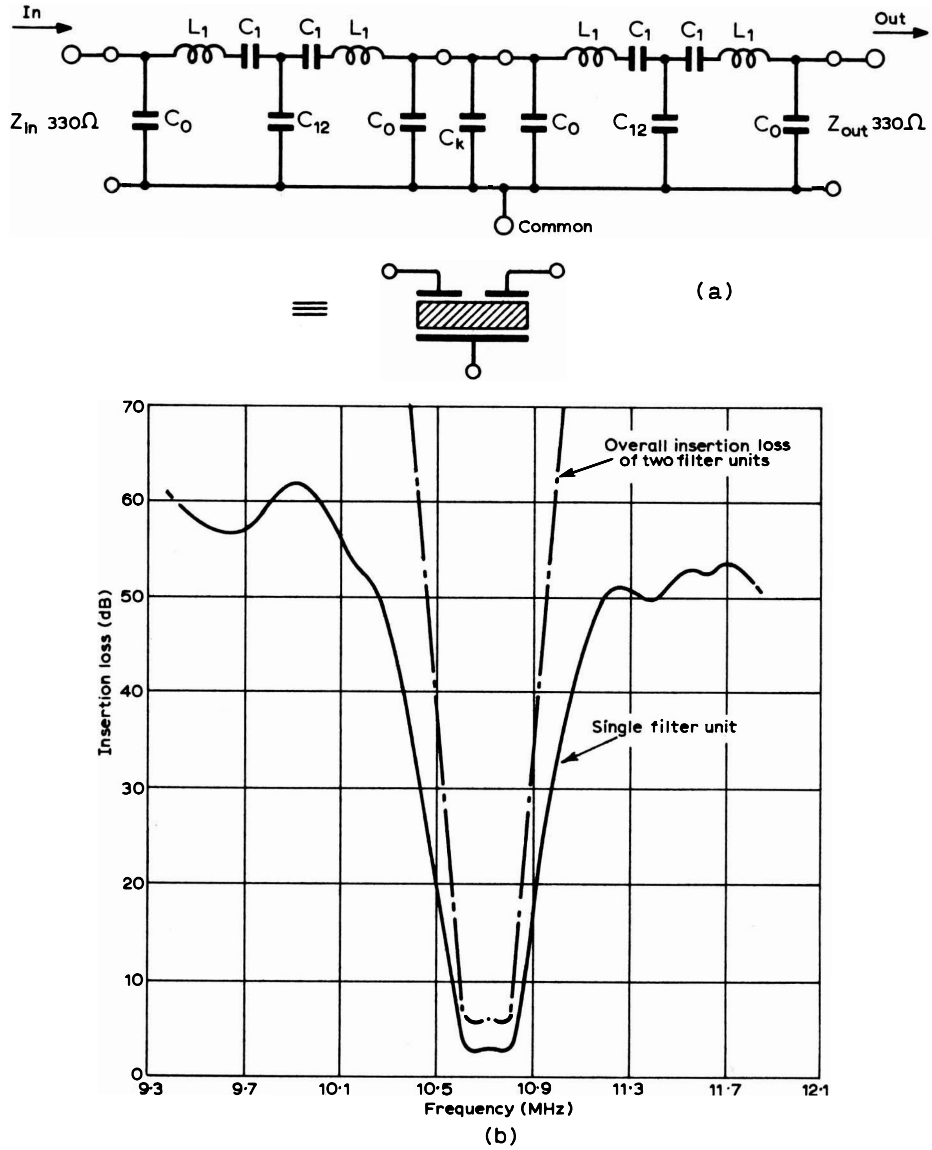


Fig. 12. Ceramic resonrtor used is equivalent to two 2-pole filters coupled by capacitor C_k (a) and has selectivity shown at (b). Two cascaded filters give a bandwidth of 220kHz at 3dB down and 560kHz at 60 dB down.

The equivalent circuit of the filter (Vernitron FM-4) is of two 2-pole filters coupled by an additional capacitor *Ck* as shown in Fig. 12. Physically all the elements are on a single ceramic substrate. The overall response is equivalent to two critically-coupled bandpass circuits in cascade. Figure 12b shows the typical selectivity of such a single unit (solid curve) consisting of a single substrate multi-pole filter with the equivalent circuit of Fig. 12a. The broken curve shows the result of using two such complete resonator units (with a suitable buffer stage between) to obtain higher selectivity. The resultant performance is more than adequate for f.m. broadcast reception, with 3-dB bandwidth of typically 220kHz and 60dB bandwidth of around 560kHz. Ripple in the pass-band is quoted as not exceeding 1dB (2dB for two stages).

Measurements confirm these figures for typical pairs of filter units in a practical amplifier. These resonators cannot be coupled directly to one another in normal use or the balance of the response curves will be upset, resulting in a highly asymmetric response—the use of driving or load impedances noticeably different from the 330-ohm design impedance will upset the degree of coupling in individual sections. This relatively low impedance of 330 ohms is perhaps one of the drawbacks of this type of ceramic resonator, although most such filters have impedances in the same region. In practice, however, the reduction in gain due to the use of such low-impedance loads in the amplifier chain is not too serious, especially as an additional low-gain buffer stage between the filter sections is needed to avoid interaction of the filters.

The most serious loss of gain from this low load impedance would occur in the mixer stage, where with a typical dual-gate f.e.t. stage as described, the voltage gain of the mixer would be reduced to a little below unity with such a load. In the tuner design the mixer is therefore modified to use a tuned load with a grounded-base buffer stage feeding a 330-ohm resistive load to which the first filter is connected. This ensures a true 330-ohm source for the filter, and results in an overall mixer gain of 24dB. The mixer load circuit is designed to work at only a moderate Q so that the tuning of this circuit is not highly critical.

Due to production tolerances the ceramic resonators are graded into frequency bands and appropriately colour coded to indicate their exact frequency tolerance. For the type used there are five groups covering a total spread of 150kHz (at 37.5-kHz intervals) around 10.7MHz. In a receiver using two such filters, both must be of the same colour group to achieve a satisfactory result. (Details of these groupings were given in the parts list.)

Variable-capacitance diodes

In the past few years some improvements have been made in the parameters of variable-capacitance diodes for tuning. These improvements have given diodes a higher Q and a wider variation in capacitance for a given voltage change. Many tuners now use such diodes exclusively for tuning the r.f. circuits, and they are becoming common in u.h.f. television tuners. The great advantage of these diodes for tuning is that the r.f. circuitry can be made very compact, thus minimizing pick-up and easing screening problems, and making circuit location independent of dial mechanism location. The main disadvantage, so far as the average constructor is concerned, is that both availability and price are at a disadvantage compared with a normal tuning capacitor at the present time.

Another common use of these diodes is a.f.c., and here the requirements are not nearly so severe as only a small variation of capacitance is required. Availability of diodes for this purpose with a smaller change of capacitance with voltage is fairly good and prices are moderate. Such a device is the Texas TIV307, whose capacitance versus voltage curve is shown in Fig. 13 (as measured by the author on three samples). This device could also be used for tuning purposes—it has just adequate capacitance variation without

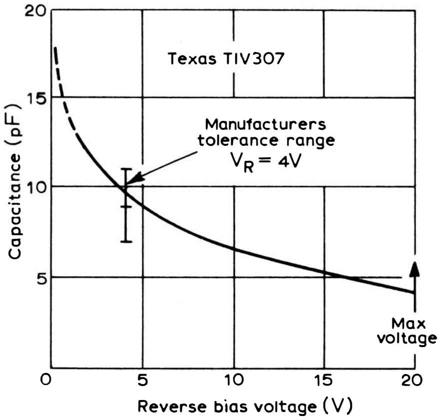


Fig. 13. Capacitance-voltage characteristic for a.f.c. diode.

using too high or too low a control voltage. Too low a control voltage is undesirable, especially in the oscillator stage, where harmonic generation and conduction of the diode become a problem at low bias voltage. Too high a voltage can be a problem either because the diode will not withstand it, or because the voltage is simply not available.

The smallest capacitance swing is in the oscillator circuit (from 87.5 + 10.7 = 98.2 to 108 + 10.7 = 118.7MHz) which has a frequency ratio f_{max}/f_{min} of 1.22, and a capacitance ratio of 1.22^2 :1 or 1.46 : 1. The r.f. circuits need a capacitance swing of $(108/87.5)^2 = 1.53:1$. A swing of from 2.5 to 7 volts would give such a change if the only capacitance were the diode. But there will always be 10-15pF of general circuit capacitance, so that a diode change from 10 to 20pF at least is necessary (or just over 2:1 variation in the diode). With some care in circuit layout and a change in L/C ratio the TIV307 could just give this swing (or its companion with higher capacitance TIV308–12pF at −4 volts), especially if a higher supply voltage than the 12 volts used in the design were available. It is the author's intention at a later date to design a diode-tuned receiver, but in the present design a normal tuning capacitor is used mainly on the grounds of cost and the difficulty of obtaining diodes in suitably matched triplets.

Alignment of tuned circuits

The tuned circuits must be aligned in reverse order, that is starting at L_5 and working back to L_1. By far the easiest way of aligning the i.f. section is to use a wobbulator centred on 10.7MHz and having a sweep frequency of 50Hz, with a peak-to-peak deviation of 1 to 2MHz. Fig. 14 shows the response of a correctly aligned i.f. amplifier and demodulator: The y-axis is the output of the tuner (1 volt/division) and the x-axis is the modulation voltage (75kHz/division). The display shown is for a moderate input, but is well into limiting. Apply the wobbulator input to L_2 via a capacitor at about 1mV level from 80 ohms. The core of L_4 is easily set for maximum gain by looking at the noise amplitude at either side of the display. As the core is moved the noise is first greatest on one side and then at the other as the resonant frequency of L_4 moves across the band. Set the core to mid-way between the positions giving maximum noise on either side.

If a wobbulator is not available then at least a signal generator must be used. Connect as for the wobbulator above to L_2 via a capacitor and apply a level of around 1mV from 80 ohms. Connect a centre-zero meter of around ±3V full scale between the output and the a.f.c. reference lead (preferably better than 10kohm/volt sensitivity). Rock the tuning of the signal generator back and forth around 10.7MHz while adjusting the core of L_5 until the positive peak excursion is equal to the negative peak excursion—Fig. 14. If the signal generator calibration is fine enough the tuning of L_5 is finally set for best linearity, plotting output voltage against frequency. If the core of L_5 is far from the correct setting, the output may be a totally positive or totally negative excursion, with no S-shape.

If a centre-zero meter is not available the 10-volt range of a multi-meter may be used connected between the output and earth. When the signal generator is far off tune the reading of the output level should be around 5.5volts (supply at 12 volts). This is equivalent to the zero centre reading using the centre-zero instrument as above. The tuning of L_5 is now set for equal deflections about 5.5 volts.

To set L_4 the signal generator is set slightly to one side of the centre frequency and the level dropped until the meter indication begins to change (i.e. drops below limiting level). Adjust L_4 to make good this change (i.e. to increase signal strength) reducing the signal generator level to keep the i.f. stages below limiting level. Continue the process until no further improvement can be made. Alternatively, the core of L_4 may be tuned for maximum noise output with no signal generator connected or with generator switched off.

While aligning the i.f. section it is a help to have the oscillator out of action to

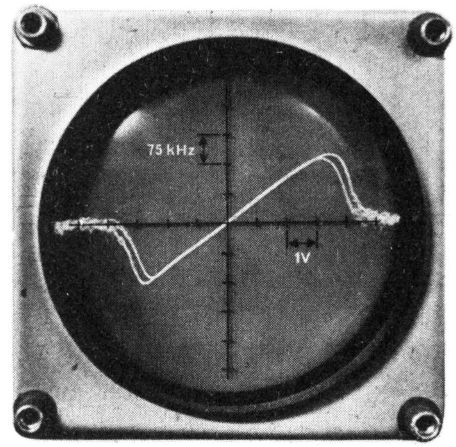


Fig. 14. Response of correctly aligned i.f. amplifier and demodulator. L_4 is set for maximum gain midway between positions giving maximum noise, either side of the display.

prevent spurious responses from i.f. harmonics. This is most easily achieved by shorting out L_3 with a single crocodile clip across the ends of the coil to connect together the end turns.

It is not sufficient to set the output level to 5.5. volts (or zero with respect to the a.f.c. reference) with an input of 10.7MHz when tuning L_5 for two reasons. First the majority of signal generators, even of very high quality, are not accurate enough to ensure a symmetrical S-shaped characteristic, and, secondly, the ceramic resonators are not necessarily peaked at 10.7MHz. If the frequency of the generator is known to within about 10kHz or better and is set to the i.f. indicated by the ceramic resonators colour code (see parts list), then L_5 may be set initially in this way. But the symmetry of the S-shaped characteristic of the detector should still be checked after setting L_4, and any slight correction made as appropriate to the core position of L_5.

Align the r.f. section in the usual way for superheterodyne receivers—set the oscillator so that the correct span of input frequencies is covered, and then adjust the r.f. circuits to track correctly.

To adjust the oscillator set the tuning capacitor to maximum capacitance and the signal generator to 87.5MHz. Adjust the variable capacitor next to the oscillator coil to receive the 87.5-MHz signal. Set the tuning capacitor to minimum capacitance and the signal generator to 108MHz. Now adjust the trimmer capacitance again noting which way this adjustment is to tune in the 108-MHz signal. (With this type of capacitor, maximum capacitance is with the silvering on the top disc towards the centre connecting pin of the capacitors three pins, and minimum is 180° from this position i.e. farthest from the middle pin.)

If the capacitance setting needs reducing at 108MHz then increase the value of L_3 by squeezing the coil to bring the turns closer together. Re-adjust the capacitor to bring the receiver back to tune at 108MHz and return to 87.5MHz and maximum capacitance of the tuning capacitor. If the trimming capacitor now needs decreasing in capacitance then the coil inductance has been increased too much.

An alternative method, possibly quicker, is to set the 108-MHz end using only the trimmer capacitor and then find to what frequency the low-end is tuned, without altering the trimmer capacitor. If this is below 87.5MHz, reduce the inductance by opening out the turns; if it is above 87.5MHz, close up the turns. Set the frequency and tuning again to 108MHz and reset the trimming capacitor. Return to the low end and again check the frequency the receiver is set to; continue this process until on reaching the low end the receiver is set to exactly 87.5MHz.

Having set the span of the oscillator, the two r.f. coils and their trimmer capacitors need adjusting to complete the alignment. This is possibly the most difficult part of the alignment procedure because of the high sensitivity of the receiver. Perhaps the simplest method is to dispense with the signal generator altogether at this point and to tune for maximum noise, with the signal generator switched off but still connected. Tracking of the r.f. coils may be set in a similar way to the oscillator coil.

Set the tuning capacitor to minimum capacitance, and tune for maximum noise using the trimmers. Set the tuning capacitor to maximum capacitance and, in the manner used for the oscillator, check whether the trimmer capacitance needs increasing or decreasing to tune for maximum noise at the low-frequency end of the dial. If the capacitance needs increasing, squeeze the coil turns closer together; if the trimmer needs decreasing in capacitance then open out the coil turns slightly. Return to the minimum value of the tuning capacitance and repeat the process from the beginning, and continue to do so until both r.f. coils need no change of tuning of the trimmer on going from one end of the dial to the other.

Further reading

The publications listed may interest those wishing to pursue various design aspects.

"MOS field-effect transistors", RCA product guide MOS160A

Data sheets on devices 40673, 3N140, 3N141, 40603, 40604 (RCA)

Data sheets on devices CA3028A, 3028B, 3053 (RCA)

"Understanding and using the dual-gate m.o.s.f.e.t." RCA application report ST-3529

"Application of dual-gate m.o.s. field-effect transistors in practical radio receivers," RCA application report ST-3486

"Integrated-circuit frequency modulation i.f. amplifiers", RCA application report ICAN-5380

"Integrated circuits for f.m. broadcast receivers", RCA application report ICAN-5269

"Use of 10.7MHz ceramic coupled-mode filters in linear i.c. i.f. strips", Vermitron application report

Data sheet on ceramic filter FM-4, bulletin 94033 (Vernitron)

Data sheet on TAA661B (SGS)

Addresses

RCA (GB) Ltd, Lincoln Way, Windmill Road, Sunbury-on-Thames, Middx

SGS Ltd, Aylesbury, Bucks

Vernitron Ltd, Thornhill, Southampton SO9 1QX, Hants

Correction

In the parts list for the tuner–published in the April issue–47-nF capacitors (10 needed) were accidentally omitted. These can be similar types to the 1-nF capacitors. In Fig. 4 a link should be added to the top left corner of L_5, between the earth area at the perimeter and the earth area under L_5. Inductor L_5 should be 10 turns–Fig. 6c. Finally, in the caption to Fig. 2, pin 2 should, of course, read pin 1.

Sixty Years Ago

May 1911. *The Marconigraph*, the original title of *Wireless World*, opened its second issue with a word of thanks to the press and the public on the reception accorded to issue No.1. The writer of a short note on the journal's content was extraordinarily prophetic when he wrote 'The other short articles and notes of recent happening in the Wireless World will no doubt prove of interest to all'.

The major technical article in the issue "A note on the Experimental Measurement of the High-frequency Resistance of Wires" was written by Dr J. A. Fleming. In essence the technique consisted of obtaining two samples of the wire and passing an a.c. current down one and a d.c. current down the other. The d.c. current was adjusted until the amount of heat produced in each wire was equal. It was then possible to calculate the r.f. resistance of the wire. The original drawing is reproduced on this page. The wires (w) are suspended in mercury (k) in air-tight glass tubes (T_1, T_2). The temperature changes affect the mercury levels and therefore the air pressure in each tube. The air bubble in paraffin oil (b) indicates the pressure difference.

This was all before the days of oscillators and the method of generating the a.c. waveform is of interest. This is how Fleming described it: 'The greatest practical difficulty is to secure a sufficiently steady high-frequency current. This was generated by employing a motor-driven alternator to give current to a large high-tension transformer raising the potential of an alternating current having a frequency of 50 to a potential of 10,000 or 20,000 volts. This voltage was used to charge one or more Leyden jars, which were discharged across a spark-gap, an air-blast on the gap being used to steady the discharge. The frequency of the oscillations so created was measured in each case by a cymometer, and the mean-square value of the current by one of the Author's hot-wire thermo-electric ammeters'.