6 HF Receivers

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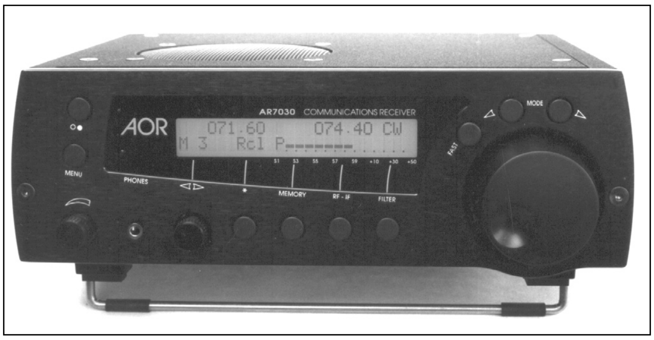


Fig 6.1: The AOR 7030 is a sophisticated receiver covering the frequency range 0-32MHz

Amateur HF operation, whether for two-way contacts or for listening to amateur transmissions, imposes stringent requirements on the receiver. The need is for a receiver that enables an experienced operator to find and hold extremely weak signals on frequency bands often crowded with much stronger signals from local stations or from the high-power broadcast stations using adjacent bands. The wanted signals may be fading repeatedly to below the external noise level, which limits the maximum usable sensitivity of HF receivers, and which will be much higher than in the VHF and UHF spectrum.

Although the receivers now used by most amateurs form part of complex, factory-built HF transceivers, the operator should understand the design parameters that determine how well or how badly they will perform in practice, and appreciate which design features contribute to basic performance as HF communications receivers, as opposed to those which may make them more user-friendly but which do not directly affect the reception of weak signals. This also applies to dedicated receivers that are factory-built, such as the one shown in Fig 6.1.

Ideally, an HF receiver should be able to provide good intelligibility from signals which may easily differ in voltage delivered from the antenna by up to 10,000 times and occasionally by up to one million times (120dB) - from less than 1µV from a weak signal to nearly 1V from a near-neighbour. **Table 1** shows the relationship between the various ways of measuring the input signals; pd (potential difference) and dBm (input power) are most commonly used.

To tune and listen to SSB or to a stable CW transmission while using a narrow-band filter, the receiver needs to have a frequency stability of within a few hertz over periods of 15 minutes or so, representing a stability of better than one part in a million. It should be capable of being tuned with great precision, either continuously or in increments of at most a few hertz.

A top-quality receiver may be required to receive transmissions on all frequencies from 1.8MHz to 30MHz (or even 50MHz) to provide 'general coverage' or only on the bands allotted to amateurs. Such a receiver may be suitable for a number of different modes of transmission - SSB, CW, AM, NBFM, data (RTTY/packet) etc - with each mode imposing different requirements in selectivity, stability and demodulation (decoding). Such a receiver would inevitably be complex and costly to buy or build.

On the other hand, a more specialised receiver covering only a limited number of bands and modes such as CW-only or CW/SSB-only, and depending for performance rather more on the skill of the operator, can be relatively simple to build at low cost.

As with other branches of electronics, the practical implementation of high-performance communications receivers has undergone a number of radical changes since their initial development in the mid-1930s, some resulting from the improved stability needed for SSB reception and others aimed at reducing costs by substituting electronic techniques in place of mechanical precision.

However, it needs to be emphasised that, in most cases, progress in one direction has tended to result in the introduction of new problems or the enhancement of others: "What we call progress is the exchange of one nuisance for another nuisance" (Havelock Ellis) or "Change is certain; progress is not" (A J P Taylor). As late as 1981, an Australian amateur was moved to write: "Solid-state technology affords commercial manufacturers cheap, large-scale production but for amateur radio receivers and transceivers of practical simplicity, valves remain incomparably superior for one-off, home-built projects."

The availability of linear integrated circuits capable of forming the heart of communications receivers combined with the increasing scarcity and hence cost of special valve types has tended to reverse this statement. It is still possible to build reasonably effective HF receivers, particularly those for limited frequency coverage, on the kitchen table with the minimum of test equipment.

for a 50 ohms power matched system	emf	pd	dBm
	1 V	0.5 V	7.0
	100 mV	50 mV	-13.0
	10 mV	5 mV	-33.0
	1 mV	0.5 mV	-53.0
possible range of signal levels	100 uV	50 uV	-80.5
	10 uV	5 uV	-93.0
	1 uV	0.5 uV	-113.0
typical receiver sensitivity			
thermal noise floor (3KHz bandwidth)	0.1 uV	50 nV	-133.0

Table 6.1: The relationship between emf, pd and dBm

Furthermore, since many newcomers will eventually acquire a factory-built transceiver but require a low-cost, stand-alone HF receiver in the interim period, the need can be met either by building a relatively simple receiver, or by acquiring, and if necessary modifying, one of the older valve-type receivers that were built in very large numbers for military communications during the second world war, or those marketed for amateur operation in the years before the virtually universal adoption of the transceiver.

Even where an amateur has no intention of building or servicing his or her own receiver, it is important that he or she should have a good understanding of the basic principles and limitations that govern the performance of all HF communications receivers.

BASIC REQUIREMENTS

The main requirements for a good HF receiver are:

- Sufficiently high sensitivity, coupled with a wide dynamic range and good linearity to allow it to cope with both the very weak and very strong signals that will appear together at the input; it should be able to do this with the minimum impairment of the signal-to-noise ratio by receiver noise, cross-modulation, blocking, intermodulation, reciprocal mixing, hum etc.
- Good selectivity to allow the selection of the required signal from among other (possibly much stronger) signals on adjacent or near-adjacent frequencies. The selectivity characteristics should 'match' the mode of transmission, so that interference susceptibility and noise bandwidth should be as close as possible to the intelligence bandwidth of the signal.
- Maximum freedom from spurious responses - that is to say signals which appear to the user to be transmitting on specific frequencies when in fact this is not the case. Such spurious responses include those arising from image responses, breakthrough of signals and harmonics of the receiver's internal oscillators.
- A high order of stability, in particular the absence of short-term frequency drift or jumping.
- Good read-out and calibration of the frequency to which the set is tuned, coupled with the ability to reset the receiver accurately and quickly to a given frequency or station.
- Means of receiving SSB and CW, normally requiring a stable beat frequency oscillator preferably in conjunction with product detection.
- Sufficient amplification to allow the reception of signals of under 1μV input; this implies a minimum voltage gain of about one million times (120dB), preferably with effective automatic gain control (AGC) to hold the audio output steady over a very wide range of input signals.
- Sturdy construction with good-quality components and with consideration given to problems of access for servicing when the inevitable occasional fault occurs.

A number of other refinements are also desirable: for example it is normal practice to provide a headphone socket on all communications receivers; it is useful to have ready provision for receiver 'muting' by an externally applied voltage to allow voice-operated, push-to-talk or CW break-in operation; an S-meter to provide immediate indication of relative signal strengths; a power take-off socket to facilitate the use of accessories; an IF signal take-off socket to allow use of external special demodulators for NBFM, FSK, DSBSC, data etc.

In recent years, significant progress has continued to be made in meeting these requirements - although we are still some way short of being able to provide them over the entire signal range of 120dB at the ideal few hertz stability. The introduction of more and more semiconductor devices into receivers has brought a number of very useful advantages, but has also paradoxically made it more difficult to achieve the highly desirable wide dynamic range. Professional users now require frequency read-out and long-term stability of an extremely high order (better than 1Hz stability is needed for some applications) and this has led to the use of frequency synthesised local oscillators and digital read-out systems; although these are effective for the purposes which led to their adoption, they are not necessarily the correct approach for amateur receivers since, unless very great care is taken, a complex frequency synthesiser not only adds significantly to the cost but may actually result in a degradation of other even more desirable characteristics.

So long as continuous tuning systems with calibrated dials were used, the mechanical aspects of a receiver remained very important; it is perhaps no accident that one of the outstanding early receivers (HRO) was largely designed by someone whose early training was that of a mechanical engineer.

It should be recognised that receivers which fall far short of ideal performance by modern standards may nevertheless still provide entirely usable results, and can often be modified to take advantage of recent techniques. Despite all the progress made in recent decades, receiver designs dating from the 'thirties and early 'forties are still capable of being put to good use, provided that the original electrical and mechanical design was sound. Similarly, the constructor may find that a simple, straightforward and low-cost receiver can give good results even when its specification is well below that now possible. It is ironical that almost all the design trends of the past 30 years have, until quite recently, impaired rather than improved the performance of receivers in the presence of strong signals!

BASIC TYPES OF RECEIVERS

Amateur HF receivers fall into one of two main categories:

(a) 'straight' regenerative and direct-conversion receivers in which the incoming signal is converted directly into audio by means of a demodulator working at the signal frequency;

(b) single- and multiple-conversion superhet receivers in which the incoming signal is first converted to one or more intermediate frequencies before being demodulated. Each type of receiver has basic advantages and disadvantages.

Regenerative Detector ('Straight' or TRF) Receivers

At one time valve receivers based on a regenerative (reaction) detector, plus one or more stages of AF amplification (ie 0-V-1, 0-V-2 etc), and sometimes one or more stages of RF amplification at signal frequency (1-V-1 etc) were widely used by amateurs. High gain can be achieved in a correctly adjusted regenerative detector when set to a degree of positive feedback just beyond that at which oscillation begins; this makes a regenerative receiver capable of receiving weak CW and SSB signals. However, this form of detector is non-linear and cannot cope well in situations where the weak signal is at all close to a strong signal; it is also inefficient as an AM detector since the gain is much reduced when the positive feedback (regeneration) is reduced below the oscillation threshold. Since the detector is non-linear, it is usually impossible to provide adequate selectivity by means of audio filters.

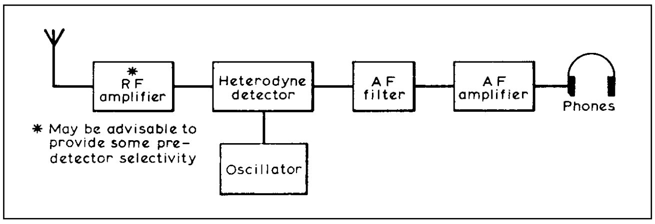


Fig 6.2: Outline of a simple direct-conversion receiver in which high selectivity can be achieved by means of audio filters

Simple Direct-conversion Receivers

A modified form of 'straight' receiver which can provide good results, even under modern conditions, becomes possible by using a linear detector which is in effect simply a frequency converter, in conjunction with a stable local oscillator set to the signal frequency (or spaced only the audio beat away from it). Provided that this stage has good linearity in respect of the signal path, it becomes possible to provide almost any desired degree of selectivity by means of audio filters (Fig 6.2).

This form of receiver (sometimes termed a homodyne) has a long history but only in the last few decades of the 20th century did it become widely used for amateur operation since it is more suited (in its simplest form) to CW and SSB reception than AM. The direct-conversion receiver may be likened to a superhet with an IF of 0kHz or alternatively to a straight receiver with a linear rather than a regenerative detector.

In a superhet receiver the incoming signal is mixed with a local oscillator signal and the intermediate frequency represents the difference between the two frequencies; thus as the two signals approach one another the IF becomes lower and lower. If this process is continued until the oscillator is at the same frequency as the incoming signal, then the output will be at audio (baseband) frequency; in effect one is using a frequency changer or translator to demodulate the signal. Because high gain cannot be achieved in a linear detector, it is necessary to provide very high AF amplification. Direct-conversion receivers can be designed to receive weak signals with good selectivity but in this form do not provide true single-sideband reception (see later); another problem often found in practice is that very strong broadcast signals (eg on 7MHz) drive the detector into non-linearity and are then demodulated directly and not affected by any setting of the local oscillator.

A crystal-controlled converter can be used in front of a direct-conversion receiver, so forming a superhet with variable IF only. Alternatively a frequency converter with a variably tuned local oscillator providing output at a fixed IF may be used in front of a direct-conversion receiver (regenerative or linear demodulator) fixed tuned to the IF output. Such a receiver is sometimes referred to as a supergainer receiver.

Two-phase and 'Third-method' Direct-Conversion Receivers

An inherent disadvantage of the simple direct-conversion receiver is that it responds equally to signals on both sides of its local oscillator frequency, and cannot reject what is termed the audio image no matter how good the audio filter characteristics; this is a serious disadvantage since it means that the selectivity of the receiver can only be made half as good as the theoretically ideal bandwidth. This problem can be overcome, though at the cost of additional complexity, by phasing techniques similar to those associated with SSB generation. Two main approaches are possible: see Fig 6.3.

Fig 6.3(a) shows the use of broad-band AF 45 degree phase-shift networks in an 'outphasing' system, and with care can result in the reduction of one sideband to the extent of 30-40dB. Another possibility is the polyphase SSB demodulator which does not require such critical component values as conventional SSB phase-shift networks.

Fig 6.3(b) shows the 'third method' (sometimes called the Weaver or Barber system) which requires the use of additional balanced mixers working at AF but eliminates the need for accurate AF phase-shift networks. The 'third method' system, particularly in its AC-coupled form [1] provides the basis for high-performance receivers at relatively low cost, although suitable designs for amateur operation are rare. Two-phase direct-conversion receivers based on two diode-ring mixers in quadrature (90° phase difference) are capable of the high performance of a good superhet.

HF Superhet Receivers

The vast majority of receivers are based on the superhet principle. By changing the incoming signals to a fixed frequency (which may be lower or higher than the incoming signals) it becomes possible to build a high-gain amplifier of controlled selectivity to a degree which would not be possible over a wide spread of signal frequencies. The main practical disadvantage with this system is that the frequency conversion process involves unwanted products which give rise to spurious responses, and much of the design process has to be concentrated on minimising the extent of these spurious responses in practical situations.

A single-conversion superhet is a receiver in which the incoming signal is converted to its intermediate frequency, amplified

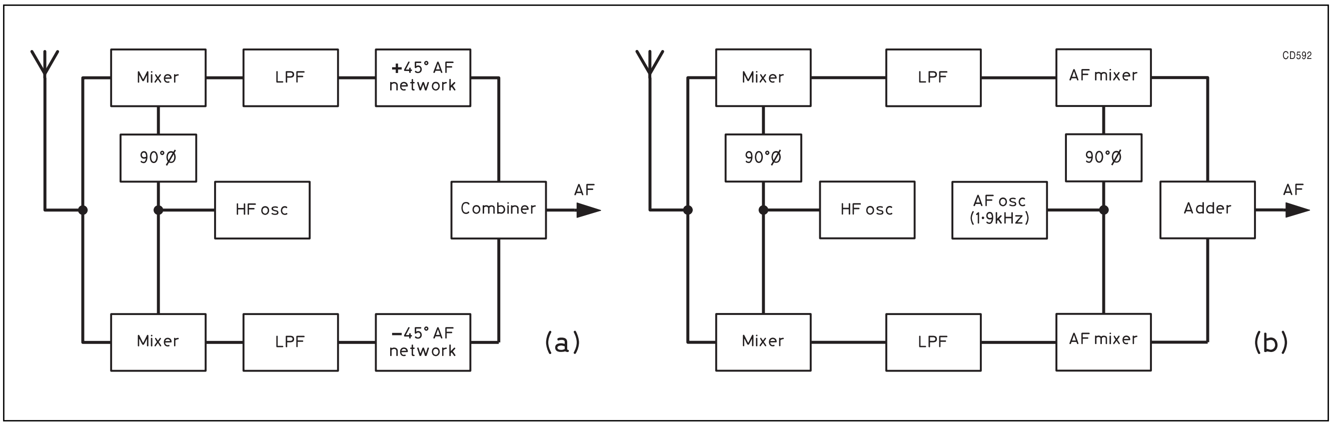
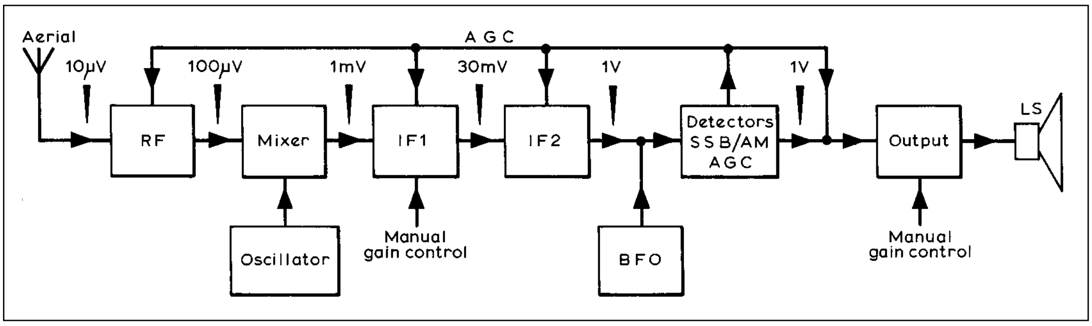


Fig 6.3: Block outline of two-phase ('autophasing') form of direct-conversion receiver. (b) Block outline of 'third method' (Weaver or Barber) SSB direct-conversion receiver

Fig 6.4: Block outline of representative single-conversion



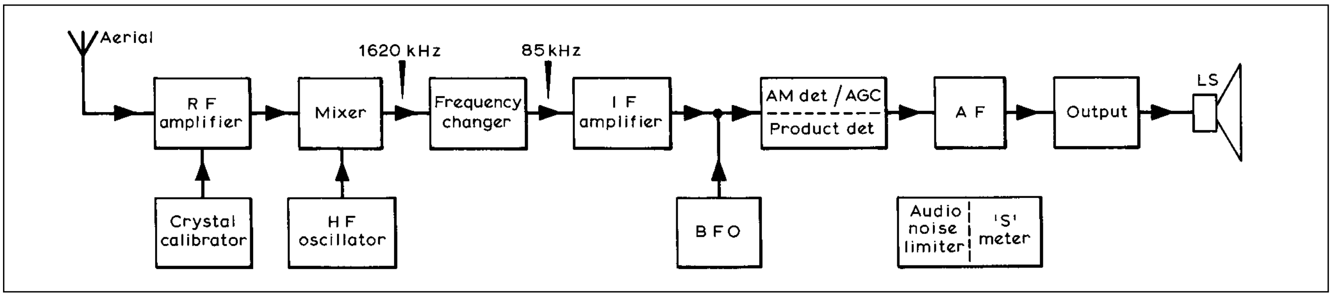


Fig 6.5: Block outline of double-conversion communications receiver with both IFs fixed

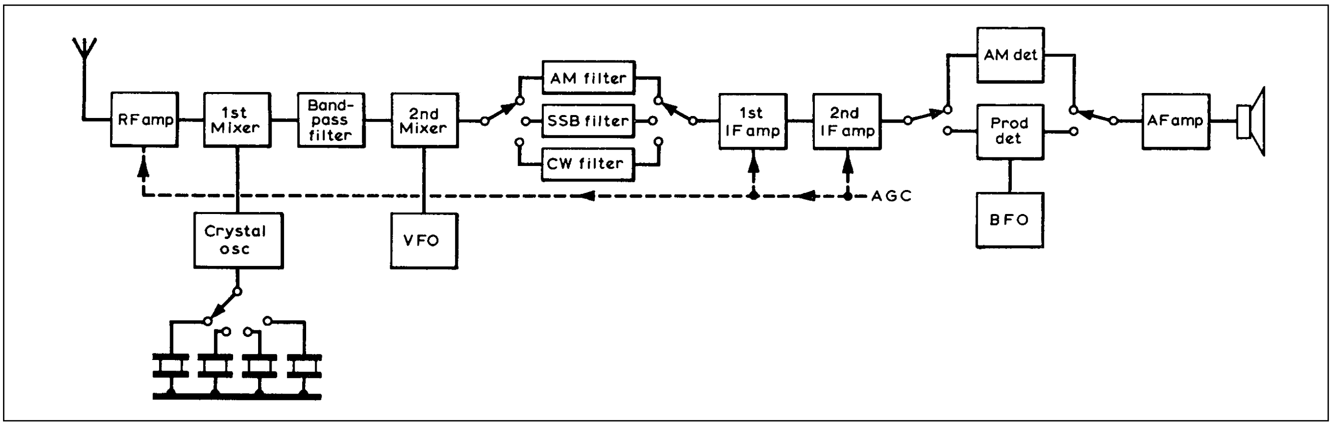


Fig 6.6: Block diagram of a double-conversion receiver with crystal-controlled first oscillator - typical of many late 20th-century designs

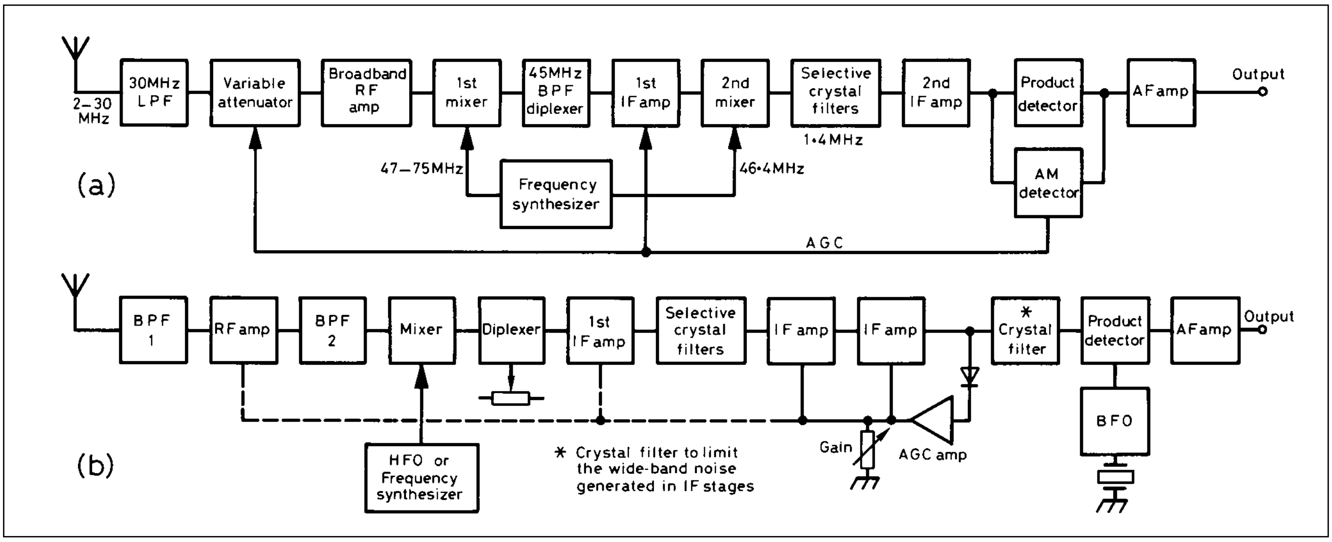


Fig 6.7: Representative architectures of modern communications receiver designs. (a) General-coverage double-conversion superhet with up-conversion to 45MHz first IF and 1.4MHz second IF. (b) Single-conversion superhet, typically for amateur bands only, with an IF in the region of 9 or 10MHz

and then demodulated at this second frequency. Virtually all domestic AM broadcast receivers use this principle, with an IF of about 455-470kHz, and a similar arrangement but with refinements was found in many communications receivers. However, for reasons that will be made clear later in this chapter, some receivers convert the incoming signal successively to several different frequencies; these may all be fixed IFs: for example the first IF might be 9MHz and the second 455kHz and possibly a third at 35kHz. Or the first IF may consist of a whole spectrum of frequencies so that the first IF is variable when tuning a given band, with a subsequent second conversion to a fixed IF. There are in fact many receivers using double or even triple conversion, and a few with even more conversions, though unless care is taken each conversion makes the receiver susceptible to more spurious responses. The block diagram of a typical single-conversion receiver is shown in **Fig 6.4**. **Fig 6.5** illustrates a double-conversion receiver with fixed IFs, while **Fig 6.6** is representative of a receiver using a variable first IF in conjunction with a crystal-controlled first local oscillator (HFO).

Many modern factory-built receivers up-convert the signal frequency to a first IF at VHF as this makes it more convenient to use a frequency synthesiser as the first HF oscillator: **Fig 6.7(a)**. As the degree of selectivity provided in a receiver increases, it reaches the stage where the receiver becomes a single-sideband receiver, although this does not mean that only SSB signals can be received.

In fact the first application of this principle was the single-signal receiver for CW reception where the selectivity is sufficient to reduce the strength of the audio image (resulting from beating the IF signal with the BFO) to an insignificant value, thus virtually at one stroke halving the apparent number of CW stations operating on the band (previously each CW signal was heard on each side of the zero beat). Similarly, double-sideband AM signals can be received on a set having a carefully controlled pass-band as though they were SSB, with the possibility of receiving either sideband should there be interference on the other. This degree of selectivity can be achieved with good IF filters or alternatively the demodulator can itself be designed to reject one or other of the sidebands, by using phasing techniques similar to those sometimes used to generate SSB signals and for two-phase direct-conversion receivers. But most receivers rely on the use of crystal or mechanical filters to provide the necessary degree of sideband selectivity, and then use heterodyne oscillators placed either side of the nominal IF to select upper or lower sidebands.

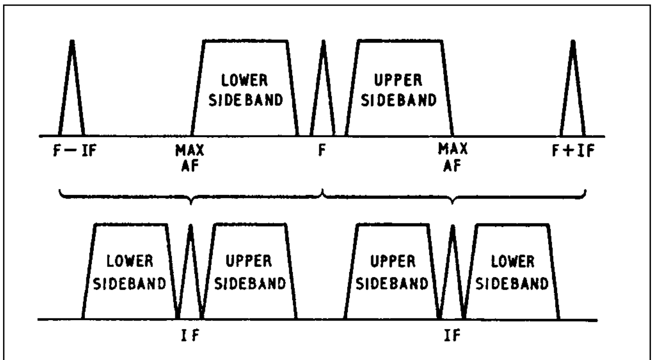


Fig 6.8: A local oscillator frequency lower than the signal frequency (ie f - IF) keeps the upper and lower sidebands of the intermediate frequency signals in their original positions. However, when the local oscillator is placed higher in frequency than the signal frequency (f + IF), the positions of the sidebands are transposed. By incorporating two oscillators, one above and the other below the input signal, sideband selection is facilitated (this is generally carried out at the final IF by switching the BFO or carrier insertion oscillator)

It is important to note that whenever frequency conversion is accomplished by beating with the incoming signal an oscillator lower in frequency than the signal frequency, the sidebands retain their original position relative to the carrier frequency, but when conversion is by means of an oscillator placed higher in frequency than the carrier, the sidebands are inverted. That is to say, an upper sideband becomes a lower sideband and vice versa (see **Fig 6.8**).

Software Defined Radio

The latest developments in professional and amateur radio communications technology are in the field of 'Software Defined Radio'. The increasing power of modern computers and digital signal processing (DSP) are bringing about unprecedented advances in radio technology and performance which looks set to soon overtake the best available from analogue designs.

Digital signal processing has been a useful technique for some years. Initially the speed of available analogue-to-digital converters (ADC) and processors limited the applicability of DSP to audio frequencies. They were nevertheless extremely useful and provided excellent audio filters, with advanced features such as noise blanking and automatic heterodyne suppression.

As devices improved at an exponential rate, it became feasible to implement the final stages of a receiver from low intermediate frequencies onwards digitally. The term Software Defined Radio is intended to convey the concept that the functionality of the radio is essentially defined by the software running on the computer which implements a significant part of the receiver's stages. This has many advantages, including the ease with which many aspects of the receiver's 'design' can altered merely by updating software.

A number of commercial designs are now available for the amateur market and have become extremely popular, such as the Flex Radio SDR1000. In many cases, these designs use a Direct Digital Synthesis (DDS) local oscillator and a mixer direct to baseband (zero IF), creating separate 90° phased I and Q signals. The computing power is provided by the operator's desktop PC, which accepts the I and Q signals from the analogue front end and demodulates them into single sideband audio. Advanced on-screen user interfaces provide the operator with numerous features such as panoramic display of a large segment of band, and incredible control over the filtering and characteristics of the receiver.Yet, the ultimate goal of SDR must be considered to be a direct digitisation of the HF Spectrum, covering 0 - 30MHz direct from the antenna, with the entire receiver implemented digitally by software. The only analogue stage in such a system would be a low-pass filter to prevent deterioration of ADC performance due to interfering signals higher than 30MHz. Until recently, the dynamic range available from ADC devices and the computing power requirements were not able to meet the demands of a true all-digital SDR.

However, the ever marching progress in semiconductor techniques and manufacture have made available high performance ADCs which have now made the all-digital HF receiver a reality. Several all-digital receivers are now available which boast high performance over the entire HF range. Whilst it may be argued that the highest performance in terms of dynamic range, sensitivity and immunity from cross-modulation is still the domain of analogue receivers, or at least analogue front ends, there is no doubt that continued advances will soon allow all-digital receivers to overtake their analogue counterparts.

For more information on SDR, please see the later chapter in this Handbook.

DESIGN TRENDS

After the 'straight' receiver, because of its relatively poor performance and lack of selectivity on AM phone signals, had fallen into disfavour in the mid-1930s, came the era of the superhet communications receiver. Most early models were single conversion designs based on an IF of 455-470kHz, with two or three IF stages, a multi-electrode triode-hexode or pentagrid mixer, sometimes but not always with a separate oscillator valve. This approach made at least one RF amplifying stage essential in order to raise the level of the incoming signal before it was applied to the relatively noisy mixer; two stages were to be preferred since this meant they could be operated in less critical conditions and provided the additional pre-mixer RF selectivity needed to reduce 'image' response on 14MHz and above. Usually a band-switched LC HF oscillator was gang-tuned so as to track with two or three signal-frequency tuned circuits, calling for fairly critical and expensive tuning and alignment systems. These receivers were often designed basically to provide full coverage on the HF band (and often also the MF band), sometimes with a second tuning control to provide electrical bandspread on amateur bands, or with provision (as on the HRO) optionally to limit coverage to amateur bands only. Selectivity depended on the use of good-quality IF transformers (sometimes with a tertiary tuned circuit) in conjunction with a single-crystal IF filter which could easily be adjusted for varying degrees of selectivity and which included a phasing control for nulling out interfering carriers.

In later years, to overcome the problem of image response with only one RF stage, there was a trend towards double- or triple-conversion receivers with a first IF of 1.6MHz or above, a second IF about 470kHz and (sometimes) a third IF about 50kHz.

With a final IF of 50kHz it was possible to provide good single-signal selectivity without the use of a crystal filter.

The need for higher stability than is usually possible with a band-switched HF oscillator and the attraction of a similar degree of band-spreading on all bands has led to the widespread adoption of an alternative form of multi-conversion superhet; in effect this provides a series of integral crystal-controlled converters in front of a superhet receiver (single or double conversion) covering only a single frequency range (for example 5000-5500kHz) This arrangement provides a fixed tuning span (in this example 500kHz) for each crystal in the HF oscillator. Since a separate crystal is needed for each band segment, most receivers of this type are designed for amateur bands only (though often with provision for the reception of a standard frequency transmission, for example on 10MHz); more recently some designs have eliminated the need for separate crystals by means of frequency synthesis, and in such cases it is economically possible to provide general coverage. The selectivity in these receivers is usually determined by a band-pass crystal filter, mechanical filter or multi-pole ceramic filter, a separate filter is being used for SSB, CW and AM reception (although for economic reasons sets may be fitted with only one filter, usually intended for SSB reception). In this system the basic 'superhet' section forms in effect a variable IF amplifier. Examples of this tunable IF architecture are the SS-R1 receiver [2] made by Squires Sanders [3], and the G2DAF design published in RadCom [4] and built by many amateurs, both in the 1960s.

In practice the variable IF type of receiver provides significantly enhanced stability and lower tuning rates on the higher frequency bands, compared with receivers using fixed IF, though it is considerably more difficult to prevent breakthrough of strong signals within the variable IF range, and to avoid altogether the appearance of 'birdies' from internal oscillators. With careful design a high standard of performance can be achieved; the use of multiple conversion (with the selective filter further from the antenna input stage) makes the system less suitable for semiconductors than for valves, particularly where broad-band circuits are employed in the front-end and in the variable IF stage.

There is now a trend back to the use of fixed IF receivers, either with single conversion or occasionally with double conversion (provided that in this case an effective roofing filter is used at the first IF). A roofing filter is a selective filter intended to reduce the number of strong signals passing down an IF chain without necessarily being of such high grade or as narrow bandwidth as the main selective filter.

To overcome the problem of image reception a much higher first IF is used; for amateur band receivers this is often 9MHz since effective SSB and CW filters at this frequency are available. This reduces (though does not eliminate) the need for pre-mixer selectivity; while the use of low-noise mixers makes it possible to reduce or eliminate RF amplification. To overcome frequency stability problems inherent in a single-conversion approach, it is possible to obtain better stability with FET oscillators than was usually possible with valves; another approach is to use mixer-VFO systems (essentially a simple form of frequency synthesis) and such systems can provide identical tuning rates on all bands, though care has to be taken to reduce to a minimum spurious injection frequencies resulting from the mixing process.

To achieve the maximum possible dynamic range, particular attention has to be given to the mixer stage, and it is an advantage to make this a balanced, or double-balanced (see later) arrangement using either double-triodes, Schottky (hot-carrier) diodes or FETs (particularly power FETs).

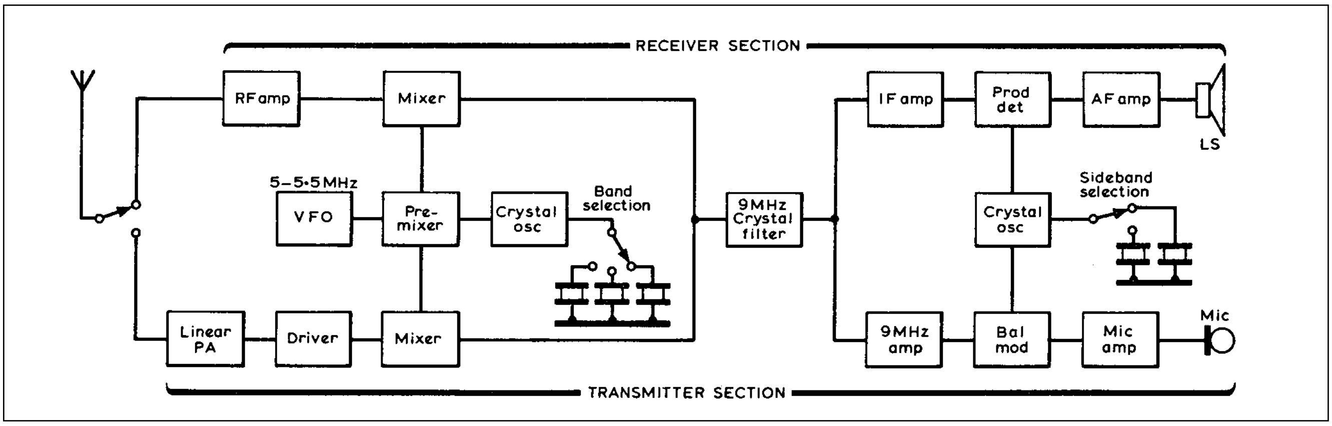


Fig 6.9: Block diagram of a typical modern SSB transceiver in which the receiver is a single-conversion superhet with 9MHz IF in conjunction with the pre-mixer form of partial frequency synthesis

A further significant reduction of spurious responses may prove possible by abandoning the superhet in favour of high-performance direct-conversion receivers (such as the Weaver or 'third-method' SSB direct-conversion arrangement); however, such designs are still only at an early stage of development.

Most modern receivers are built in the form of compact transceivers functioning both as receiver and transmitter, and with some stages common to both functions (**Fig 6.9**). Modern transceivers use semiconductor devices throughout. Dual-gate FET devices are generally found in the signal path of the receiver. Most transceivers have a common SSB filter for receive and transmit; this may be a mechanical or crystal filter at about 455kHz but current models more often use crystal filters at about 3180, 5200, 9000kHz or 10.7MHz, since the use of a higher frequency reduces the total number of frequency conversions necessary.

One of the fundamental benefits of a transceiver is that it provides common tuning of the receiver and transmitter so that both are always 'netted' to the same frequency. It remains, however, an operational advantage to be able to tune the receiver a few kilohertz around the transmit frequency and vice versa, and provision for this incremental tuning is often incorporated; alternatively many transceivers offer two oscillators so that the transmit and receive frequencies may be separated when required.

The most critical aspect of modern receivers is the signal-handling capabilities of the early (front-end) stages. Various circuit techniques are available to enhance such characteristics: for example the use of balanced (push-pull) rather than single-ended signal frequency amplifiers; the use of balanced or double-balanced mixer stages; the provision of manual or AGC-actuated antenna-input attenuators; and careful attention to the question of gain distribution.

An important advantage of modern techniques such as linear integrated circuits and wide-band fixed-tuned filters rather than tuneable resonant circuits is that they make it possible to build satisfactory receivers without the time-consuming and constructional complexity formerly associated with high-performance receivers. Nevertheless a multiband receiver must still be regarded as a project requiring considerable skill and patience.

The widespread adoption of frequency synthesisers as the local HF oscillator has led to a basic change in the design of most factory-built receivers, although low-cost synthesisers may not be the best approach for home-construction. Such synthesisers cannot readily (except under microprocessor control) be 'ganged' to band-switched signal-frequency tuned circuits; additionally, mechanically-ganged variable tuning of band-switched signal-frequency and local oscillator tuned circuits as found in older communications receivers would today be a relatively high-cost technique.

These considerations have led to widespread adoption in factory-built receivers and transceivers of 'single-span' up-conversion multiple-conversion superhets with a first IF in the VHF range, up to about 90MHz, followed by further conversions to lower IFs at which the main selectivity filter(s) are located.

In such designs, pre-selection before the first mixer or preamplifier (often arranged to be optionally switched out of circuit) may simply take the form of a low-pass filter (cut off at 30MHz) or a single wide-band filter covering the entire HF band. Higher-performance receivers usually fit a series of sub-octave band-pass filters, with electronic switching (preferably with PIN diodes).

With fixed filtering, even of the sub-octave type, very strong HF broadcast transmissions will be present at the mixer(s) and throughout the 'front-end' up to the main selectivity filter(s). To enable weak signals to be received free of intermodulation products, this places stringent requirements on the linearity of the front-end. The use of relatively noisy low-cost PLL frequency synthesisers also raises the problem of 'reciprocal mixing' (see later). For home-construction of high-performance receivers, the earlier design approaches are still attractive, including the 'old-fashioned' concept of achieving good pre-mixer selectivity with high-Q tuned circuits using variable capacitors rather than electronic tuning diodes. Diode switching rather than mechanical switching can also significantly degrade the intermodulation performance of receivers. Further, it should be noted that reed relay switching can often introduce sufficient series resistance to seriously degrade the Q of tuned circuits unless reeds especially selected for their RF properties are used.

DIGITAL TECHNIQUES

The availability of general-purpose, low-cost digital integrated circuit devices made a significant impact on the design of communication receivers although, until the later introduction of digital signal processing, their application was primarily for operator convenience and their use for stable, low-cost frequency synthesisers rather than their use in the signal path.

By incorporating a digital frequency counter or by operation directly from a frequency synthesiser, it is now normal practice to display the frequency to which the receiver (or transceiver) is tuned directly on matrices of light-emitting diodes or liquid crystal displays. This requires that the display is offset by the IF from the actual output of the frequency synthesiser or free-running local oscillator. Such displays have virtually replaced the use of calibrated tuning dials.

Frequency synthesisers are commonly 'tuned' by a rotary shaft-encoded switch which can have the 'feel' of mechanical capacitor tuning of a VFO, but this may be supplemented by push-buttons which enable the wanted frequency to be punched

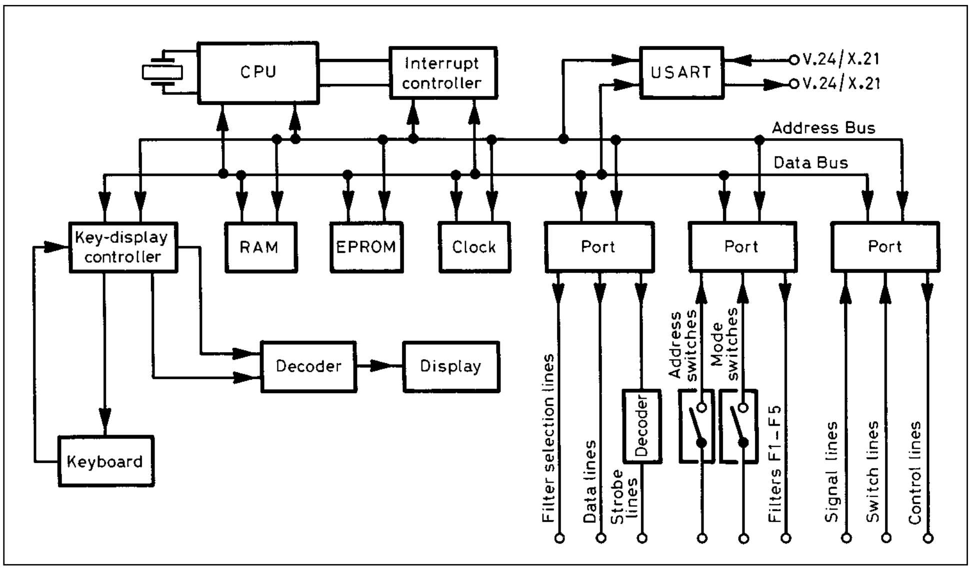


Fig 6.10: Architecture of the elaborate internal computer system found in a modern professional fully-synthesised receiver (DJ2LR)

in. Common practice is for the frequency change per knob revolution to be governed by the rate at which the knob is spun, to speed up large changes of frequency. With a synthesiser the frequency may change in steps of 100Hz, 10Hz or even 1Hz, and it is desirable that this should be free of clicks and should appear to be almost instantaneous. Many of the factory-built equipments incorporate digital memory chips which can be programmed with frequencies to which it is desired to return to frequently. It should be noted that a phase-locked oscillator has an inherent jitter that appears as phase and amplitude noise that can give rise to reciprocal mixing and an apparent raising of the noise floor of the receiver. Digital direct frequency synthesis (DDS) can reduce phase noise and is being increasingly used.

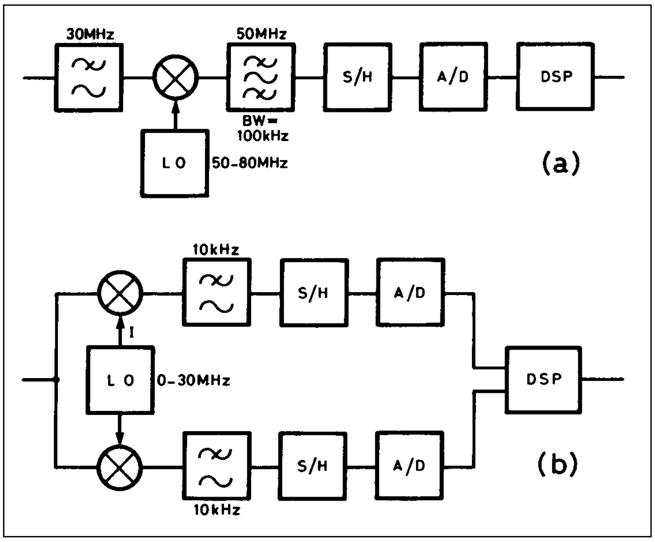


Fig 6.11: Basic arrangements for hybrid analogue/digital receivers incorporating digital signal processing. (a) Single-channel, single-conversion superhet (in practice a double-conversion analogue front-end is more likely to be employed to bring the IF signal down to 1.6 or 3MHz). (b) Dual-channel homodyne approach: S/H, sample and hold; A/D, analogue/digital converter; DSP, digital signal processing

Digital techniques may be used to stabilise an existing free-running VFO by continuously 'sampling' the frequency over pre-determined timing periods and then applying a DC correction to a varactor forming part of the VFO tuned circuit. The timing periods can be derived from a stable crystal oscillator and the technique is capable of holding a reasonably good VFO to within a few hertz. Here again some care is needed to prevent the digital pulses, with harmonics extending into the VHF range, from affecting reception.

Microprocessor control of user interfaces can include driving the tuning display, memory management (including BFO and pass-band tuning), frequency and channel scanning; data bus to RS-232 conversion. As stressed by Dr Ulrich Rohde, DJ2LR, several key points need to be observed:

- Keypad and tuning-knob scanning must not generate any switching noises that can reach the signal path.
- All possible combination of functions such as frequency steps, operating modes, BFO frequency offset and pass-band tuning should be freely and independently programmable and storable as one data string in memory.
- It is desirable that multilevel menus should be provided for easy use and display of all functions. This includes not only modes (USB, LSB, CW etc) but also AGC attack and decay times. Such parameters should be freely accessible and independently selectable.

It should be understood that most of the above microprocessor functions are for user convenience and do not add to the basic performance (apart from frequency stability) of a receiver; and unless care is exercised may in practice degrade performance. **Fig 6.10** shows an example of the computer control architecture of a modern receiver.

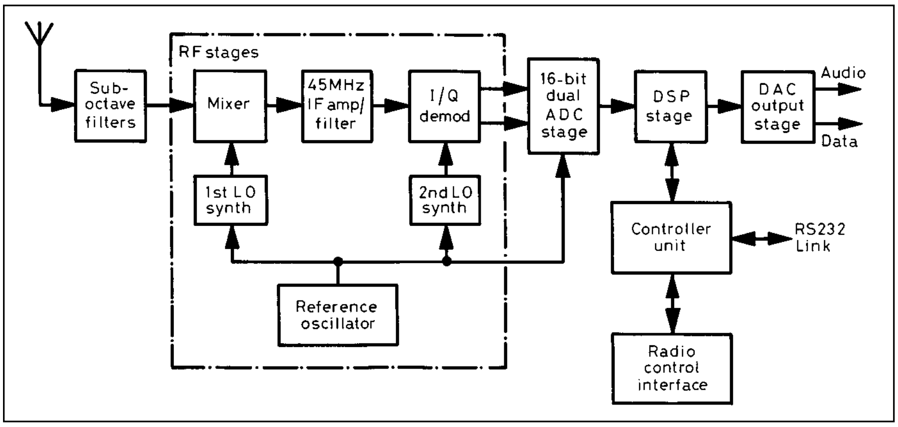


Fig 6.12: Outline of prototype high-performance analogue / digital communications receiver with baseband digitisation following two-phase I/Q demodulation (Roke Manor Research Ltd)

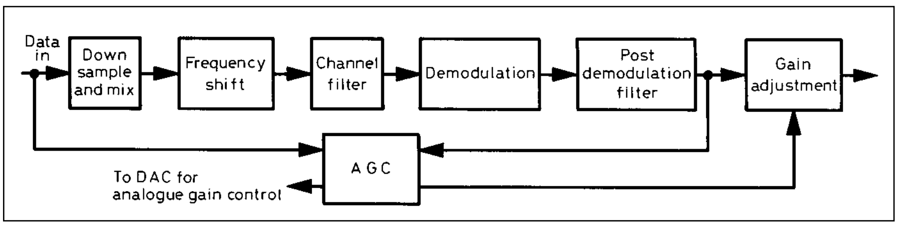


Fig 6.13: Software structure for the STC receiver

Digital Signal Processing

Digital techniques have been used in HF receivers for several decades, notably in the form of frequency synthesisers, digital frequency displays, frequency memories, microprocessor controlled mode-switching, variable tuning rates etc. By the mid-1980s, professional designers were engaged in further extensions in the use of digital rather than analogue approaches in the main signal path of HF receivers.

Hybrid analogue/digital receivers incorporating DSP, with digitisation of incoming signals at audio frequency (baseband frequency) or at a relatively low IF have been available for some years. Advantages of DSP include the ability to provide accurately shaped selectivity characteristics at a lower cost than a full complement of analogue (crystal) band-pass filters as fitted in high-grade, multimode professional receivers. Furthermore, digital technology offers a lower component count, easier factory assembly, higher reliability and the lower costs that can ensure from using standard digital devices. New features become possible in such areas as AGC, noise cancellation etc.

Professional hybrid analogue-digital HF receivers (see **Figs 6.11-6.13**) have been implemented in three main configurations:

- Two-phase (SSB demodulation) direct-conversion with baseband DSP filtering.
- Superhet front-end with 'zero-IF' (super-gainer type) with two-phase SSB demodulation and baseband DSP.
- Multi-conversion superhet with a low final IF (LF or MF), with digitisation at the low IF of up to about 500kHz (**Fig 6.14**). This third approach may use 'undersampling' below the Nyquist rate (Nyquist's theory requires that sampling should be at least twice the maximum frequency of the sampled waveform) with the aliasing products eliminated by means of relatively low-grade analogue filtering (eg ceramic filter). For example, a sampling frequency of 96kHz may be used to digitise signals at an IF of 456kHz.

Amateur-radio transceivers have been marketed with DSP filtering as an optional extra, in this case at baseband frequencies behind a conventional SSB front-end with a product detector after analogue SSB filters.

DSP at other than audio frequency has the problem that the dynamic range is adversely affected by the need that much of the receiver gain must be ahead of the analogue to digital converter (ADC), and therefore must be under automatic gain control. With effective AGC the dynamic range may be specified as better than 120dB but it should be appreciated that the instantaneous dynamic range will be much lower than this, since current ADCs cannot cope with large dynamic ranges. It should be appreciated that a large instantaneous dynamic range is highly desirable when trying to copy a weak narrow-band signal immediately alongside a very strong signal. The linearity and resolution of the ADC is thus a key requirement for a high-performance receiver. It should be realised that if a 1 microvolt input signal is to be digitised such that a 20dB SINAD ratio (Signal plus Noise plus Distortion to Noise plus Distortion) is required, the digitisation of the wanted signal needs to be to 4 bits. Handling a 0.25V signal off-tune then demands an ADC with a resolution of 24 bits per sample if the ADC is to have a linear transfer characteristic (ie equal voltage steps per bit).

There is no doubt, however, that DSP can provide excellent filter characteristics programmed to match the required bandwidth of different modes. For example, an SSB filter's attenuation at a 500Hz offset from the upper band edge can be over 60dB without any adjustments required and is insensitive to

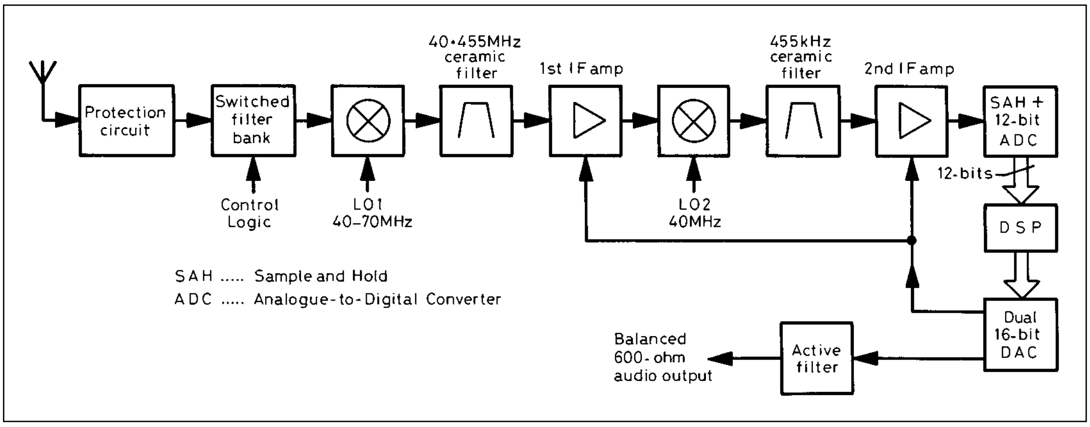


Fig 6.14: Block diagram of the STC marine HF band analogue / digital receiver with sub-Nyquist sampling of 455kHz IF

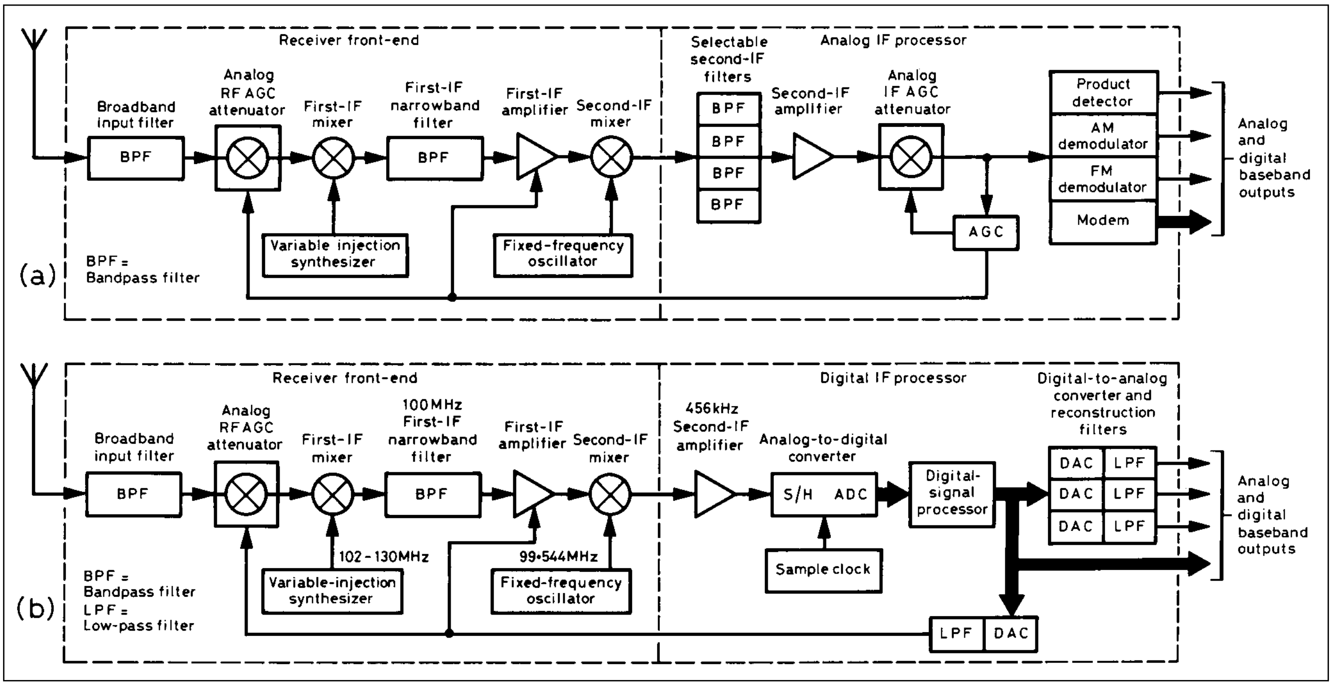


Fig 6.15: (a) Outline of typical professional analogue communications receiver with several band-pass filters and demodulation circuits for the different modes. (b) Use of a software-reconfigurable digital signal processor at the 455kHz second IF potentially reduces cost by eliminating the multiple band-pass filters etc

external component and environmental changes, and without the phase delays that can affect performance of data transmissions.

The same DSP filter can be programmed to provide a series of band-pass filters (eg 500Hz, 1kHz, 2.7kHz, 6kHz etc), each having shape factors superior to a complete set of high-cost crystal filters. (**Fig 6.15**) In practice, many of the advantages of DSP filtering may be lost due to limitations in the analogue front-end and in the ADC as noted above.

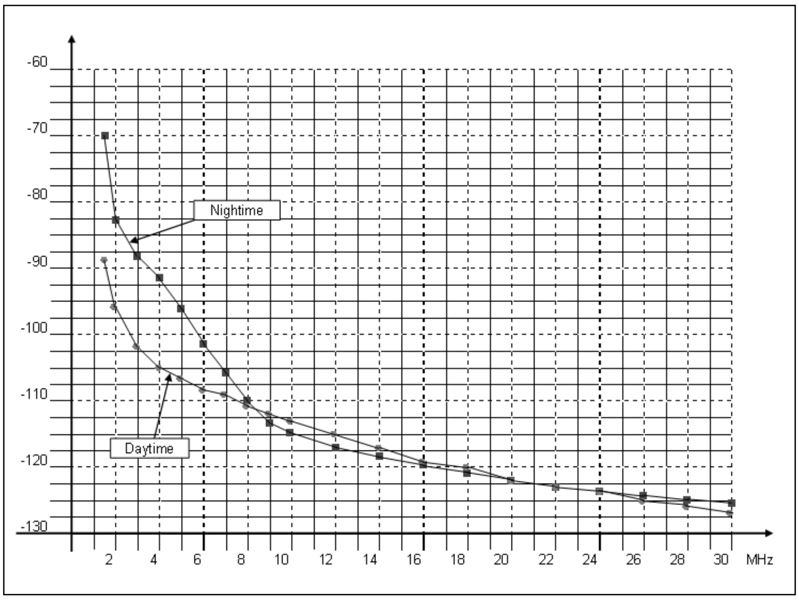


FIg 6.16: Background noise (dBm) in a rural location from a half-wave dipole and in a 3kHz bandwidth

RECEIVER SPECIFICATION

The performance of a communications receiver is normally specified by manufacturers or given in equipment reviews, or stated in the various constructional articles. It is important to understand what these specifications mean and how they relate to practical requirements in order to know what to look for in a good receiver. It will be necessary to study any specifications with some caution, since a manufacturer or designer will usually wish to present his receiver in the most favourable light, and either omit unfavourable characteristics or specify them in obscure terms. The specifications do not tell the whole story: the operational 'feel' may be as important as the electrical performance; the 'touch' of the tuning control, the absence of mechanical backlash or other irregularities, the convenient placing of controls, the positive or uncertain action of the band-change switch and so on will all be vitally important.

Furthermore, there is a big difference between receiver measurements made under laboratory conditions, with only locally generated signals applied to the input, and the actual conditions under which it will be used, with literally hundreds of amateur, commercial and broadcasting signals being delivered by the antenna in the presence of electrical interference and possibly including one or more 'blockbusting' signals from a nearby transmitter.

Sensitivity

Weak signals clearly need to be amplified more than strong ones in order to provide a satisfactory output to the loudspeaker, headphones or data modem. However, there are limits to this process set by the noise generated within the receiver and the external noise picked up by the antenna. What is important to the operator is the signal-to-noise ratio (SNR) of the output signal and how this compares with the SNR of the signal delivered by the antenna. Ideally these would be the same, in which case the noise factor (F) would be unity.

Noise generated within the receiver is most important when it arises in those parts of the receiver where the incoming signal is still weak, ie in the early (front-end) stages; this noise is, within limits, under the control of the designer.

External noise includes atmospheric noise, which is dependent on both frequency, time of day and ionospheric conditions, and also on local man-made electromagnetic noise, which will usually be more significant in urban locations than in rural sites due to the multiplicity of electrical and electronic appliances. **Fig 6.16** shows the results of background noise measurements made in a quiet rural location across the HF spectrum. It illustrates the high level of background noise at the low frequency end and the difference between the daytime and night-time levels

Noise power is usually regarded as distributed evenly over the receiver's bandwidth so that the effective noise level is reduced if the receiver is operated at the minimum bandwidth appropriate for the mode in use (eg about 2.1 or 2.7kHz for SSB, about 300Hz for CW).

Noise Figure and Noise Factor

Noise figure defines the maximum sensitivity of a receiver without regard to its pass-band (bandwidth) or its input impedance, and will be determined in the front-end of the receiver. It is defined as the ratio:

$$\text{Noise factor } F = \text{SNRin/SNRout}$$

where the SNR is noise power ratio. When expressed logarithmically:

$$\text{Noise Figure NF} = 10 \log_{10} (\text{SNRin/SNRout}) \text{ dB}$$

The minimum equivalent input noise for a receiver at room temperature (290K) is -174dBm/Hz. This is given by:

$$P = kTB$$

where P is the power in watts, k is Boltzmann's constant (1.38 x 10^{-23} J/K) and B is the bandwidth in Hertz. The noise floor of the receiver is then given by:

$$\text{Noise floor} = -174 + \text{NF} + 10 \log_{10}\text{B dBm}$$

where NF is the noise figure in decibels, dBm is the level in decibels relative to 1 milliwatt and, for an input resistance of 50Ω, this means that -107dBm is 1 microvolt, and B is the receiver noise bandwidth in Hertz.

For a CW or SSB receiver, this is approximately the 6dB bandwidth of the narrowest filter: for FM the 6dB bandwidth of the narrowest IF filter can be used as a reasonable approximation. However, the SNR out of an FM detector displays a non-linear relationship to the input SNR, depending upon the detector and IF amplifier type.

So a receiver with a 3kHz bandwidth and a 10dB noise figure will have an equivalent input noise floor of-130dBm. That is to say, if the receiver was perfect, it would behave as if the input noise was at -130dBm. The relationship between sensitivity and the various ways of describing noise is shown in **Fig 6.17**.

Because of galactic, atmospheric and man-made noise always present on HF there is little need for a receiver noise figure of less than 15-17dB on bands up to about 18-20MHz, or less than about 10dB up to 30MHz, even in quiet sites. It may, however, be an advantage if the first stages (preamplifier or

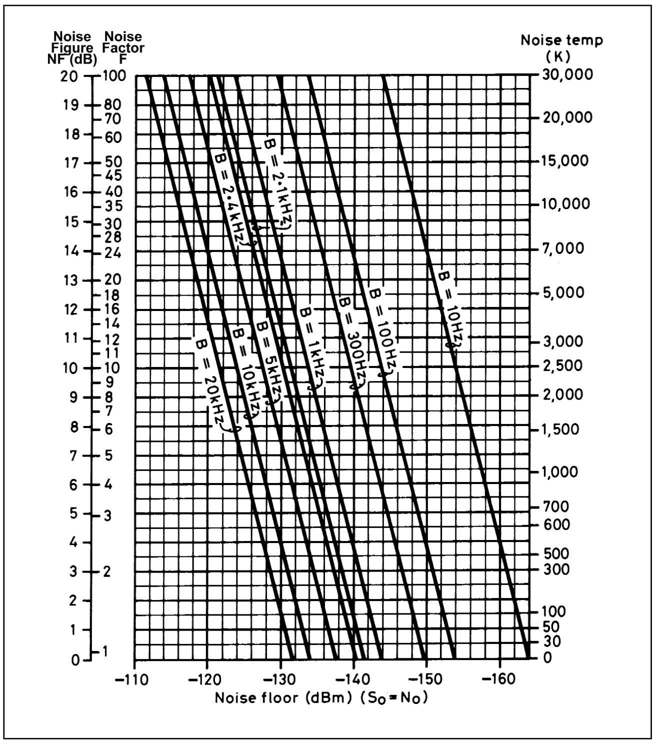


Fig 6.17: Relationship between the sensitivity of a receiver as defined in terms of noise factor, noise figure (dB) or noise temperature and the noise floor (noise = signal) in dBm for various receiver bandwidths. Note how the minimum detectable signal reduces with narrower bandwidths for the same noise factor

mixer or post-mixer) have a lower noise figure since this will permit good reception with an electrically short antenna or allow the use of a narrow-band filter, which attenuates the signal power even if providing an impedance step-up, to be used between antenna and the mixer stage (ie improved pre-mixer selectivity).

It should be noted that excessive sensitivity is likely to impair the strong-signal handling capabilities of the receiver.

Signal-to-noise ratio (or, as measured in practice, more accurately signal-plus-noise to noise ratio) gives the minimum antenna input voltage to the receiver needed to give a stated output SNR with a specified noise bandwidth. The input voltage for a given input power depends on the input impedance of the receiver; for modern sets this is invariably 50 or 75Ω but for older (though still useful) receivers this may be 400Ω and such sets may thus appear wrongly to be less sensitive unless the impedance is taken into account.

The signal delivered by the antenna from a weak incoming signal may well be of the order of -130dBm, representing 0.14μV across a matched 50Ω line. It should be noted that where sensitivity is defined in terms of SNR, this depends not only on input impedance but also on the output SNR which, while usually 10dB, may occasionally be 6dB; it will also depend as noted above on the noise bandwidth of the receiver.

It is useful to note also the minimum discernible signal (MDS) given in dBm, which is defined as where the signal is equal to the noise voltage. In practice MDS is measured by determination of signal having a very small difference from the noise floor, such as a 1dB SNR. This introduces very little error in the measurement of MDS. The MDS is thus the noise floor of the receiver and represents the power required from a signal generator to produce a 3dB (S+N)/N output, ie Where the input signal equals the background noise.

MDS (dBm) = -174dBm + 10 log B + NF

where B is the receiver noise bandwidth in Hz and NF is the Noise Figure in decibels.

The noise floor of HF receivers is conventionally given in -dBm in a 3kHz noise bandwidth and in practice may range from about -120dBm to about -145dBm. Alternatively, receiver noise in a 3kHz bandwidth at 50Ω impedance may be represented by an equivalent signal expressed in dbmV EMF, which will equal the noise figure less 26dB.

With modern forms of mixers, it is possible to achieve a noise figure better than 10dB so that no preamplification is required to achieve maximum usable sensitivity on HF up to about 21MHz; however, low-gain amplification may still be advisable in order to minimise radiation of the local oscillator output from the antenna and to permit the use of narrow-band or resonant input filtering. Oscillator radiation is limited for amateur equipment to a maximum level of -57dBm conducted to the antenna for emissions below 1GHz, and to -47dBm above 1GHz, by the European Radiocommunications Committee Recommendation on Spurious Emissions.

The effect of cascading stages is given by the well-known equation due to Friis:

$$F_t = F_1 + [(F_2 - 1)/g_1] + [(F_3 - 1)/g_1 g_2] + \ldots$$

where (all in power ratios and NOT decibels) F_t is the overall noise factor, F_2, F_3 are the individual stage noise figures and g_1, g_2 are the individual stage gains. To convert back to Noise Figure in dB we take the $\log_{10}$ and multiply by 10 ie 10 $\log_{10}$ Ft. Obviously, the series can be extended to cover as many stages as necessary.

For design work, this can usefully be incorporated into a spreadsheet approach for rapid evaluation of changes; however, the values must be converted from decibels to absolute power ratios.

Selectivity

The ability of a receiver to separate stations on closely adjacent frequencies is determined by its selectivity. The limit to usable selectivity is governed by the bandwidth of the type of signal which is being received.

For high-fidelity reception of a double-sideband AM signal the response of a receiver would need ideally to extend some 15kHz either side of the carrier frequency, equivalent to 30kHz bandwidth; any reduction of bandwidth would cause some loss of the information being transmitted. In practice, for average MF broadcast reception the figure is reduced to about 9kHz or even less; for communications-quality speech in a double-sideband

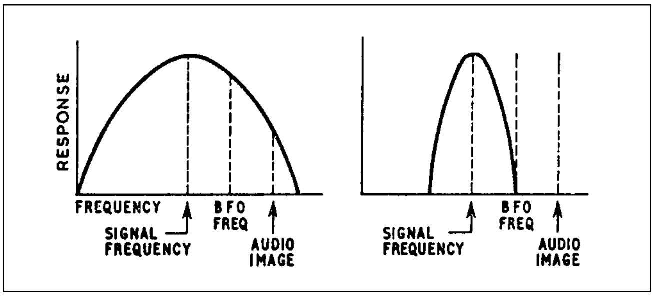


Fig 6.18: How a really selective receiver provides 'single-signal' reception of CW. The broad selectivity of the response curve on the left is unable to provide substantial rejection of the audio 'image' frequency whereas with the more selective curve the audio image is inaudible and CW signals are received only on one side of zero-beat

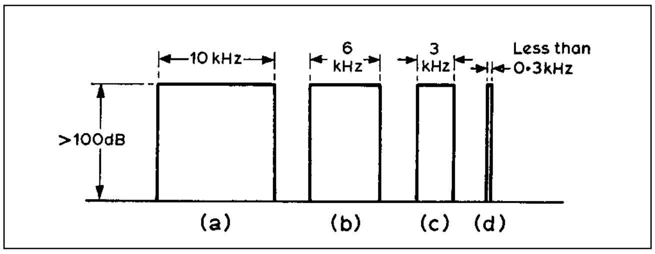


Fig 6.19: The ideal characteristics of the overall band-pass of a receiver are affected by the type of signals to be received. (a) This would be suitable for normal broadcast reception (DSB signals) permitting AF response to 5kHz. (b) Suitable for AM phone (AF to 3kHz). (c) For SSB the band-pass can be halved without affecting the AF response (in the example shown this would be about 300 to 3300Hz). (d) Extremely narrow channels (under 100Hz) are occupied by manually keyed CW signals but some allowance must usually be made for receiver or transmitter drift and a 300Hz bandwidth is typical - by selection of the carrier insertion oscillator frequency any desired AF beat note can be produced

system we require a bandwidth of about 6kHz; for single-sideband speech about half this figure or 3kHz is adequate, and filters with a nose bandwidth of 2.7 or 2.1kHz are used, providing, in the case of a 2.7kHz filter, audio frequencies from 300 to 3000Hz with little loss of intelligibility. For CW, at manual keying speeds, the minimum theoretically possible bandwidth will reduce with speed from about 100Hz to about 10Hz for very slow Morse. Fig 6.18 shows how 'single signal' reception is achieved with a narrow filter. Excessively narrow filters with good shape factors make searching difficult - and many operators like to have some idea of signals within a few hundred hertz of the wanted signal. Ideally, again, we would like to receive just the right bandwidth, with the response of the receiver then dropping right off as shown in Fig 6.19, to keep the noise bandwidth to a minimum. Although modern filters can approach this response quite closely, in practice the response will not drop away as sharply over as many decibels as the ideal.

To compare the selectivity of different receivers, or the same receiver for different modes, a series of curves of the type shown in Fig 6.20 may be used. There are two ways in which these curves should be considered: first the bandwidth at the nose, representing the bandwidth over which a signal will be received with little loss of strength; the other figure - in practice every bit as important - is the bandwidth over which a powerful signal is still audible, termed the skirt bandwidth.

The nose bandwidth is usually measured for a reduction of not more than 6dB, the skirt bandwidth for a reduction of one thousand times on its strength when correctly tuned in, that is 60dB down. These two figures can then be related by what is termed the shape factor, representing the bandwidth at the skirt divided by the bandwidth at the nose.

The idealised curves of Fig 6.19, which have the same bandwidth regardless of signal strength, would represent a shape factor of 1; such a receiver cannot be designed at the present state of the art, although it can be approached by digital filters and some SSB filters; the narrower CW filters, although much sharper at the nose, tend to broaden out to about the same bandwidth as the SSB filter and thus have a rather worse shape factor, although this may not be a handicap. Typically a high-grade modern

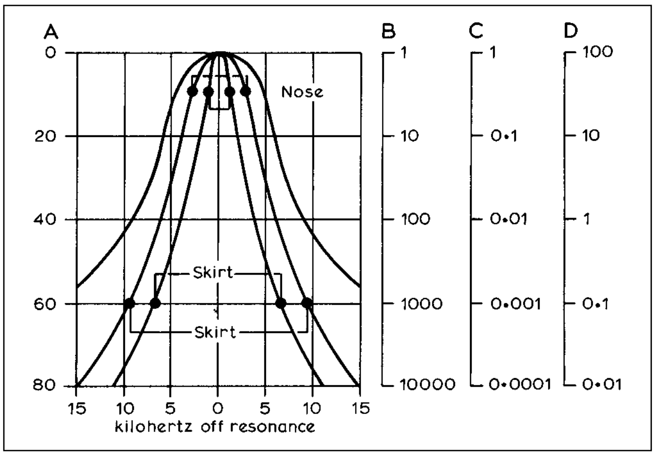


Fig 6.20: The ideal vertical sides of Fig 6.19 cannot be achieved in practice. The curves shown here are typical. These three curves represent the overall selectivity of receivers varying from the 'just adequate' broadcast curve of a superhet receiver having about four tuned IF circuits on 470kHz to those of a moderately good communications receiver. A, B, C and D indicate four different scales often used to indicate similar results. A is a scale based on the attenuation in decibels from maximum response; B represents the relative signal outputs for a constant output; C is the output voltage compared with that at maximum response; D is the response expressed as a percentage

receiver might have an SSB shape factor of 1.2 to 2 with a skirt bandwidth of less than 5kHz. Note that, although 6 and 60dB are conventionally used for nose and skirt, some radios are specified at 3 and 30dB, or even 6 and 30dB - it is important to be aware of what figures are used when comparing one filter or receiver with another.

Fig 6.21 shows both the specification and the measured response of an LF-D6 ceramic filter indicating a shape factor of 1.14.

It should be noted that such specifications are determined when applying only one signal to the input of the receiver; unfortunately in practice this does not mean that a receiver will be unaffected by very strong signals operating many kilohertz away from the required signal and outside the IF pass-band; this important point will be considered later in this chapter.

It therefore needs to be stressed that the effective selectivity cannot be considered solely in terms of static characteristics determined when just one test signal is applied to the input, but rather in the real-life situation of hundreds of signals present at the input: in other words it is the dynamic selectivity which largely determines the operational value of an amateur HF receiver.

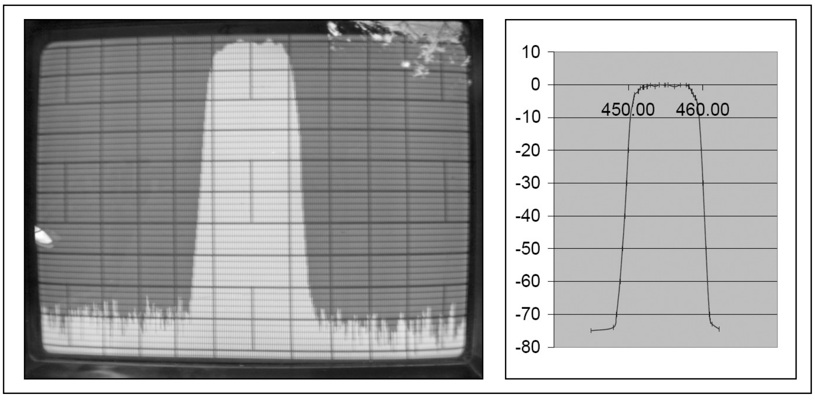


Fig 6.21: (left) Measured and (right) theoretical selectivity of a LF-D6 ceramic filter

With the advent of block IF amplifiers, it has become common practice to put one high quality filter in front of the IF gain block, but this does nothing to limit the amount of noise generated within the block from reaching the detectors. Before gain blocks, selectivity was provided progressively along the amplifier as part of the interstage coupling, 'distributed selectivity' in today's parlance. Ideally, some additional selectivity is required prior to the detectors to eliminate this additional source of IF noise.

Strong-signal Performance

The ability of a receiver to receive effectively weak signals in the presence of much stronger signals at frequencies not far removed from the desired signal is today recognised as more important than extreme sensitivity. The strong signals may come from local or medium-distance amateur stations (particularly during contest operation) or at any time from high-power (typically 500kW) broadcast stations in bands adjacent to the 7, 10, 14, 18 and 21MHz amateur bands. Strong signals can result in desensitising (blocking) the receiver, cross-modulation and/or the generation of spurious intermodulation products.

Cross-modulation, Blocking & Intermodulation

Even with a receiver that is highly selective to one signal down to the -60dB level, there remains the problem of coping with numbers of extremely strong signals. When an unwanted signal is transmitting on a frequency that is well outside the IF passband of the receiver, it may unfortunately still affect reception as a result of cross-modulation, blocking or intermodulation.

When any active device such as a transistor or valve is operated with an input signal that is large enough to drive the device into a non-linear part of its transfer characteristic (ie so that some parts of the input waveform are distorted and amplified to different degrees) the device acts as a 'modulator', impressing on the wanted signal the modulation of the strong signal by the normal process of mixing. When a very strong signal reaches a receiver, the broad selectivity of the signal-frequency tuned circuits (and often any tuned circuits, including IF circuits, prior to the main selective filter) means it will be amplified, along with the wanted signal, until one or more stages are likely to be driven into a non-linear condition. It should be noted that the strong signal may be many kilohertz away from the wanted signal, but once this cross-modulation has occurred there is no means of separating the wanted and unwanted modulation. A strong CW carrier can reduce the amplification of the wanted signal by a similar process which is in this case called desensitisation, or in extreme cases blocking.

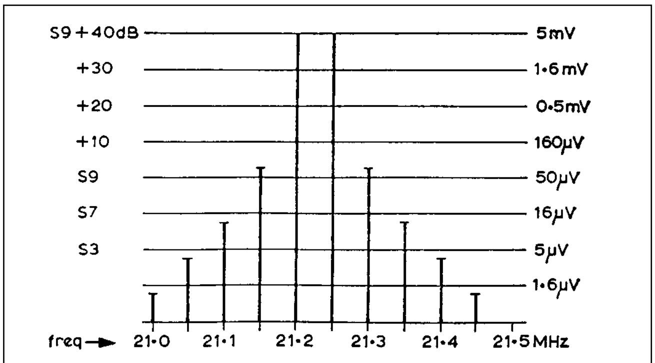


Fig 6.22: Intermodulation products. This diagram shows the effect of two very strong signals on 21,200kHz and 21,250kHz reaching the front-end of a typical modern transceiver and producing spurious signals at 50kHz intervals. Note the S9 signals produced as the third-order products f1 + (f1 - f2) and f2 - (f1 - f2). Three very strong signals will produce far more spurious signals, and so on

In these processes there need be no special frequency relationship between wanted and unwanted signals. However, a further condition arises when there are specific frequency relationships between wanted and unwanted signals, or between two strong unwanted signals; a process called intermodulation: see Fig 6.22. Intermodulation is closely allied to the normal mixing process but with strong signals providing the equivalent of local oscillators, unwanted intermodulation products (IPs) can result from many different combinations of input signal. Fig 6.22 shows the effects of two signals intermodulating; it may be shown that as far as the products of any order are concerned, there are 2n(n - 1) intermodulation products actually capable of producing signals on the desired frequency. Using a sub-octave filter ahead of the intermodulating stage will reduce this by a factor of two. Products of the form $f_1 + f_2 - f_3 = f_{tune}$ must also be considered. Using a sub-octave or less bandwidth filter will lead to there being a total number of intermodulation products:

$$T = n(n - 1) + 0.5n(n - 1)(n - 2)$$

where T is the total number of intermodulation products and n is the total number of input signals.

Because of the potentially large number of products, the effect is to raise the effective noise floor of the receiver. This occurs in a manner analogous to white light being created from a mixture of all colours (frequencies): similarly, noise is produced from a mixture of a large number of frequencies of random amplitude and phase distribution.

It should be appreciated that cross-modulation, blocking and intermodulation products can all result from the presence of extremely strong unwanted signals applied to any stage having insufficient linearity over the full required dynamic range. The solution to this problem is either to reduce the strength of unwanted signals applied to the stage, or alternatively to improve the dynamic range of the stage.

Clearly the more amplifying stages there are in a receiver before the circuits or filters which determine its final selectivity, the greater are the chances that one or more may be overloaded unless particular care is paid to the gain distribution in the receiver (see later). From this it follows that multiple-conversion receivers are more prone to these problems than single-conversion or direct-conversion receivers.

It is often difficult for an amateur to assess accurately the performance of a receiver in this respect; undoubtedly many multi-conversion superhets fall far below the desired performance. Even high-grade receivers are likely to be affected by S9 + 60dB signals 50kHz or more away from the wanted signal.

The susceptibility of receivers to these forms of interference depends on a number of factors: notably the type of active devices used in the front-end of the receiver; the phase noise performance of the various oscillators, the pattern of gain distribution through the receiver and how this is modified by the action of AGC. On receivers suffering from this problem (often most apparent on 7MHz where weak amateur stations may be sought alongside extremely strong broadcast stations), considerable improvement often results from the inclusion of an attenuator in the antenna feeder, since this will reduce the unwanted signals to an extent where they may not cause spurii before reducing the wanted signal to the level where it cannot be copied.

If sufficient dynamic range could be achieved in all stages before the final selectivity filter, then selectivity could be determined at any point in the receiver (for example at first, second or third IF or even at AF).

In practice, some early semiconductor receivers were unable to cope with undesired signals more than about 20-30dB stronger than a required signal of moderate strength, though their 'static' selectivity was extremely good. The overall effect of limited dynamic range depends upon the actual situation and band on which the receiver is used; it is of most importance on 7MHz or where there is another amateur within a few hundred yards. Even where there are no local amateurs, an analysis of commercial HF signals has shown that typically, out of a total of some 3800 signals logged between 3 and 29MHz at strengths more than 10dB above atmospheric noise, 154 were between 60-70dBμV, 72 between 70-80dBμV, 36 between 80-90dBμV and 34 between 90-100dBμV. If weak signals are to be received satisfactorily the receiver needs to be able to cope with signals some 100dB stronger (ie up to about S9 + 50dB). This underlines the importance of achieving front-end stages of extremely wide dynamic range or alternatively providing sufficient pre-mixer selectivity to cut down the strength of all unwanted signals reaching the mixer (preferably to well under 100mV). It has been shown that even the use of a small antenna such as a 3/4 G5RV at 8m (25ft) can lead to a substantial intermodulation problem with receivers having only a 0dBm third-order intercept point.[5]

The effects arise from non-linearity in the receiver and can be specified in terms of the receiver's dynamic range, although it is important to note that the dynamic range needs to be specified separately for blocking, cross-modulation, intermodulation and reciprocal mixing (see later). The most useful and most critical specification is given by the third-order intercept point or as the spurious-free dynamic range.

Strong-signal performance and a wide dynamic range have become widely recognised as important and highly desirable characteristics of receiver performance, although it can be argued that, at least for the Amateur Service (particularly CW), accepted methods of laboratory measurement may not provide a realistic specification. It is conventional practice to make dynamic range measurements using two signal generators with frequencies spaced 20kHz apart. However, where receivers incorporate relatively low-cost frequency synthesisers, the spacing between the two signals may have to be increased to 50 or even 100kHz in order to overcome the effects of synthesiser phase noise (made even more significant by the trend to up-conversion to a VHF intermediate frequency because VHF oscillators are generally worse than HF ones in this respect). It would be more useful if instantaneous dynamic range measurements were made with frequencies spaced 2 or 5kHz apart. It should also be remembered that in practice, the receiver will be required to cope not with just two locally generated signals but with dozens of strong broadcast carriers which will reach the vulnerable mixer stage unless filtered out (or at least reduced) by pre-mixer selectivity. Multiple carriers produce a multitude of IMPs which may resemble noise, thus raising the noise floor and decreasing the sensitivity of the receiver.

In specifying receiver dynamic range use is made of the concept of intercept points, although it is important to realise that these are purely graphic presentations based on two-signal laboratory measurements and do not represent directly the signal-handling properties of a receiver. Third-order intermodulation products are measured in the laboratory with the aid of a spectrum analyser and two high-performance signal generators. The intercept point is found with the aid of a graph with axes representing output power in dBm plotted against input power in dBm, on which the wanted signal and the N-order intermodulation product outputs are extended to the points where the IMPs intercept the wanted signal plot. This will occur at signal inputs well above those that can in practice be handled by the receiver

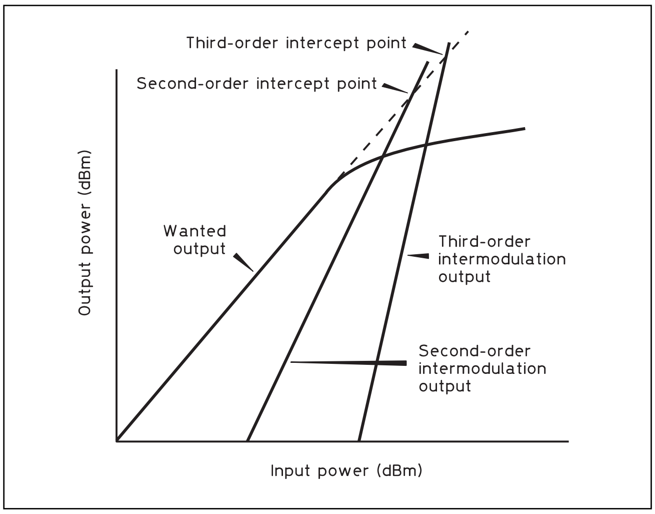


Fig 6.23: Graphical representation of receiver dynamic range performance

(Fig 6.23). Note that there are two intercept points which may be quoted - the input intercept point, and the output intercept point. These differ by the gain of the device, it not always being specified as to which is being used. Generally, commercial considerations lead to whichever is the biggest being the one quoted.

With both axes logarithmic, the second-order IMPs will have twice, and the third-order IMPs approximately three times, the slope of a wanted signal. For third-order IMPs, N = 3 and the IMP power is approximately proportional to the cube of the input powers. In other words, for every 1dB increase in the input powers, the IP3 output power increases by approximately 3dB. When plotted graphically there will thus be a point at which the IP3 output crosses and overtakes the plot of a wanted signal.

This shows that with very strong off-frequency signals, IMPs would completely block out the wanted signal. But for specification purposes we are more concerned to know the point at which the IMPs rise above the noise floor and can just be heard as unwanted interference.

A useful guide to the signal-handling performance of a receiver is given by the spurious-free dynamic range (SFDR). The upper end is defined by the signal level applied to the receiver input which produces third-order IMD products equal to the noise floor of the receiver under test. The lower end is defined as the minimum input signal at the noise floor of the receiver, ie the minimum discernible signal (MDS).

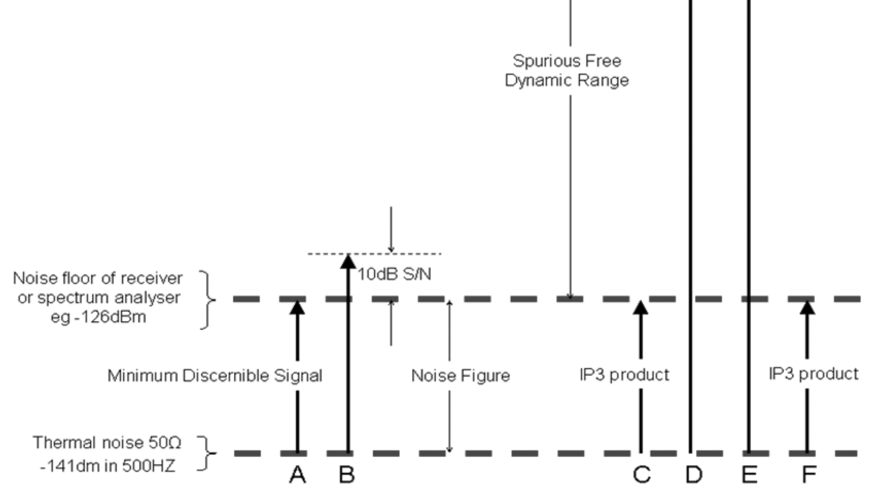
The SFDR is given by 2/3 x (IP3 - No) where SFDR is in decibels, IP3 is the third-order intercept point in dBm and No is the receiver noise floor in dBm. In the laboratory the receiver noise floor will be that of the internally generated noise and this will usually be significantly below the noise floor with the receiver connected to an antenna. The full SFDR may thus not be usable in practice, and it may be more useful to know the maximum input signal level, Pi(max) that produces third-order IMD products equal to the receiver noise floor:

$$P_{i(max)} = 1/3\,(2IP + N_o)$$

where Pi(max) is the maximum input signal in dBm, IP is the third-order intercept point in dBm and No the receiver noise floor in dBm.

SFDR is sometimes termed the two-tone dynamic range, but note that receiver dynamic range may be defined differently, for example as the blocking dynamic range (BDR) representing the difference between the 1dB compression point and the noise floor. With double-balanced mixers, a rule of thumb puts the third-order intercept point some 10 to 15dB above the 1dB compression point but, since the spurious-free IMD range is restricted to where the IMPs rise above the noise floor, the blocking dynamic range will be significantly greater than the SFDR. For example, an extremely-high-performance front-end with a third-order input intercept point of +33dBm, a 1dB input compression point of +14.3dBm and a minimum discernible signal of -133.4dBm would have a spurious-free dynamic range of 111dB but a blocking dynamic range of 147.7dB. In practice most receivers would have performance specifications considerably below this example. However, as will be shown later, for these levels of performance to be of any use, the reciprocal mixing performance must also be considered. In this respect, the phase noise limited dynamic range (PNDR) is as important as the SFDR, and the two should be approximately equal.

Fig 6.24: Relationship between the significant receiver parameters



In practice, the noise levels at the receiver input are quite likely to be high enough to provide a bottom end limit to sensitivity, and certain unpublished measurements suggest that a SFDR of about 95 to 100dB, used in conjunction with a switched 10 or 20dB antenna attenuator is able to provide a more than adequate performance. The use of an antenna attenuator may well be considered as an admission of defeat: it should NOT be considered as a substitute for anything less than 95dB of SFDR.

Merely because crystal and mechanical filters are passive devices, they should not be considered as non-contributors to the intermodulation performance of the receiver. Crystal filters are notorious for their IM contribution: the higher the frequency, the worse the performance. One cause for this is the mechanical stress in the crystal for a given voltage across it increases with frequency because of the decrease in crystal thickness, and when the limit of Hook's law is approached, intermodulation occurs. Another factor is concerned with finishing of the crystal, where the final lapping can have marked effects. It is worth while turning a filter round, as it may well perform far better one way round than the other. Mechanical filters of the older variety are also quite poor, although modern filters have performances comparable to crystal filters - input intermodulation intercept points of +16 to +30dBm being achieved.

Generally speaking, cross-modulation is not a common effect in modern receivers: the effects of intermodulation and reciprocal mixing are more prevalent. On FM, of course, cross-modulation as such does not occur.

It needs to be stressed that in order to achieve state-of-the-art meaningful dynamic performance of a receiver, extreme care must be taken not only with the vulnerable mixer stages but throughout the receiver, up to and including the final selectivity filters. On the other hand, budget-conscious constructors/purchasers can take heart from the fact that receivers with far less stringent specifications may still prove entirely adequate for many forms of amateur activity not involving the reception of very weak signals under the most hostile conditions such as international HF contests, or, for example, 7MHz night-time operation in the presence of strong broadcast carriers.

Fig 6.24 shows the relationship between the various receiver parameters.

Automatic Gain Control (AGC)

The fading characteristics of HF signals and the absence of a carrier wave with SSB make the provision of effective AGC an important characteristic of a communications receiver, although it should be stressed that no AGC may be preferable to a poor AGC system that can degrade overall performance of the receiver. Unfortunately, unwanted dynamic effects of an AGC system cannot be deduced from the usual form of receiver specification which usually provides only limited information on the operation of the AGC circuits, indicating only the change of audio output for a specified change of RF input. For example, the specification may state that there will be a 3dB rise in audio output for an RF signal input change from 1mV to 50mV. This represents a high standard of control, provided that the sensitivity of the receiver is not similarly reduced by strong off-tune signals, and that the control acts smoothly throughout its range without introducing intermodulation distortion etc.

Basically, AGC is applied to a receiver to maintain the level of the wanted signal output at a more or less constant value, while ensuring that none of the stages is overloaded, with consequent production of IMPs etc. The control voltages, derived usually at the end of the IF amplifying stages, are applied to a number of stages in the signal path while usually ensuring that the IF gain is reduced first, and the RF/first mixer later, in order to preserve the SNR.

When AGC is applied to an amplifier, this shifts the operating point and may affect both its dynamic range and the production of intermodulation distortion. One partial solution is the use of an AGC-controlled RF attenuator(s) ahead of the first stage(s).

All AGC systems are designed with an inherent delay based on resistance-capacitance time-constants in their response to changes in the incoming signal; too-rapid response would result in the receiver following the audio envelope or impulsive noise peaks. The delays are specified as the attack time, ie the time taken for the AGC to act, and the decay time for which it continues to act in the absence or fade of the wanted signal (Fig 6.25).

For AM signals, now only rarely encountered in amateur radio, the envelope detector may be used directly to generate the AGC voltage. In this case the attack and delay times need to be fast enough to allow the AGC system to respond to fading, but slow enough not to respond to noise pulses or the modulation of the carrier. Typically time-constants are about 0.1 to 0.2s.

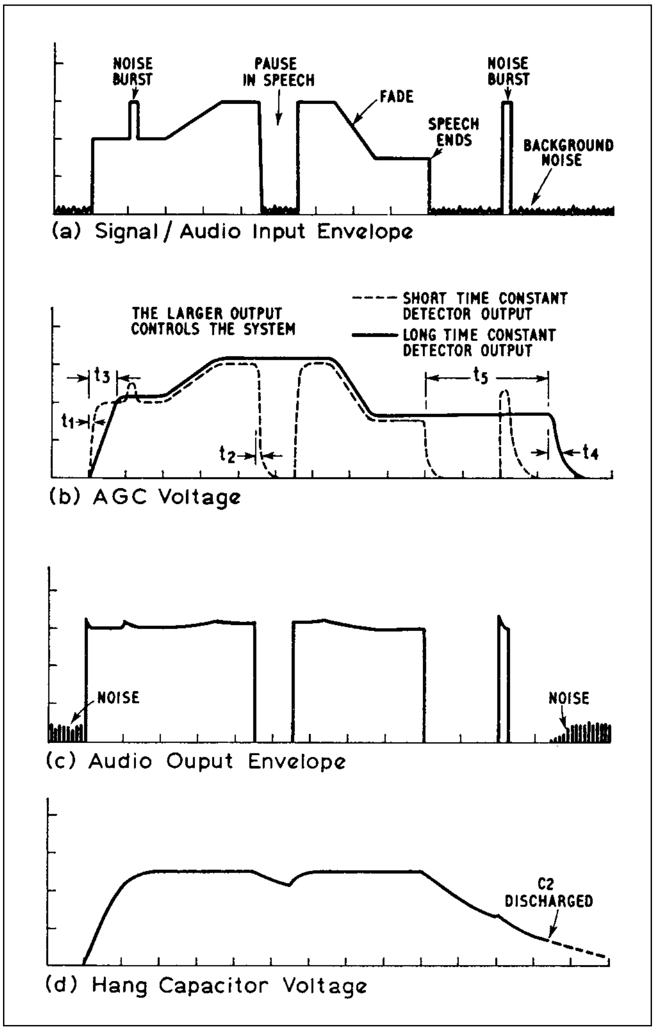


Fig 6.25: The behaviour of dual time-constant AGC circuits under various operating conditions, where t_1 is the fast detector rise time, t_2 the fast detector decay time, t_3 the slow detector rise time, t_4 the slow detector fall time, and t_5 the hang time

For amateur SSB signals there is no carrier level and the AGC voltage must be derived from the peak signal level. This is sometimes done by using the AF signal from a product detector but it is more satisfactory to incorporate a dedicated envelope detector to which a portion of the final IF signal is fed. This avoids the effects of the AGC signal disappearing at low audio beat notes. For SSB, a fast attack time is needed but the release (decay) time needs to be slow in order not to respond to the brief pauses of a speech transmission. Such a system is termed hang AGC and can be used also for CW reception. Typically the attack time needs to be less than 20ms and the release time some 200 to 1000ms; with a receiver intended also for AM reception it is useful to have a shorter release time (say 25ms) available for fast AGC. Hang AGC systems are often based on two time-constants to allow the system to be relatively unaffected by noise pulses while retaining fast attack and hang characteristics.

For both SSB and CW reception, it is important that the BFO or carrier insertion oscillator should not affect the AGC system and for many years this problem was sidestepped by turning the AGC system off during CW reception. Even with modern AGC systems, it may be useful to be able to turn the system off, particularly where final narrow-band CW selectivity depends on post-demodulation filtering, with the result that the AGC system will react to signals within the bandwidth of the final IF system which may be about 3kHz unless narrow-band IF crystal filters are fitted.

The distribution of gain control is an important, but often neglected function. Obviously, gain could be controlled by an antenna attenuator, but the result would be that the signal-to-noise ratio was that obtained for the weakest input signal. Ideally, an increase of input signal of 20dB would provide an increase in SINAD of 20dB, but there is usually an ultimate receiver SINAD limit of 40 to 50dB. A test of SINAD improvement ratio is usefully but rarely done: a signal is fed into the receiver to provide a 20dB SINAD ratio, and is increased in 10dB steps. A reasonable receiver will show about 28 to 29dB SINAD for the first step, and no less than 37dB SINAD for the second.

Oscillator Noise and Reciprocal Mixing

A single-conversion superhet requires a variable HF oscillator which should not only be stable but should provide an extremely 'pure' signal with the minimum of noise sidebands. Any variation of the oscillator output in terms of frequency drift or sudden 'jumps' will cause the receiver to detune from the incoming signal. The need for a spectrally pure output is less readily grasped, yet it is this feature which often represents a practical limitation on the performance of modern receivers. This is due to noise sidebands or jitter in the oscillator output. Noise voltages, well known in amplifiers, occur also in oscillators, producing output voltages spread over a wide frequency band and rising rapidly immediately adjacent to the wanted oscillator output.

The noise jitter and sidebands immediately adjacent to the oscillator frequency are particularly important. When a large interfering signal reaches the mixer on an immediately adjacent channel to the wanted signal, this signal will mix with the tiny noise sidebands of the oscillator (the sidebands represent in effect a spread of oscillator frequencies) and so may produce output in the IF pass-band of the receiver: this effect is termed reciprocal mixing. Such a receiver will appear noisy, and the effect is usually confused with a high noise factor.

At HF the 'noise' output of an oscillator falls away very sharply either side of the oscillator frequency, yet it is now recognised that this noise may be sufficient to limit the performance of a receiver.

Particular care is necessary with some forms of frequency synthesisers, including those based on the phase-locking of a free-running oscillator, since these can often produce significantly more noise sidebands and jitter than that from a free-running oscillator alone.

Synthesisers involving a number of mixing processes may easily have a noise spectrum 40 to 50dB higher than a basic LC oscillator. A tightly controlled phase-locked oscillator with variable divider might be some 20 to 30dB higher than an LC oscillator, but possibly less than this where the VCO is inherently very stable and needs only infrequent 'correction'.

For LC and crystal oscillators, FETs are often recommended. However, suitable bipolar transistors operated in common base (where the a cut-off frequency is highest) with a 27 ohm unbypassed emitter resistor have been shown to be capable of equal results. The use of fairly high supply voltages and currents (increasing the tank circuit power) allows high SNRs to be obtained, albeit at the price of power consumption and possible thermal effects. In VCOs, the use of two varactor diodes back-to-back to avoid conduction at any point of the RF cycle is also recommended.

Although 'hyperabrupt' diodes offer the greatest tuning range, they are often of lower Q than the graded junction types, and it is in any case best to have the lowest possible sensitivity in megahertz per volt that can be used.

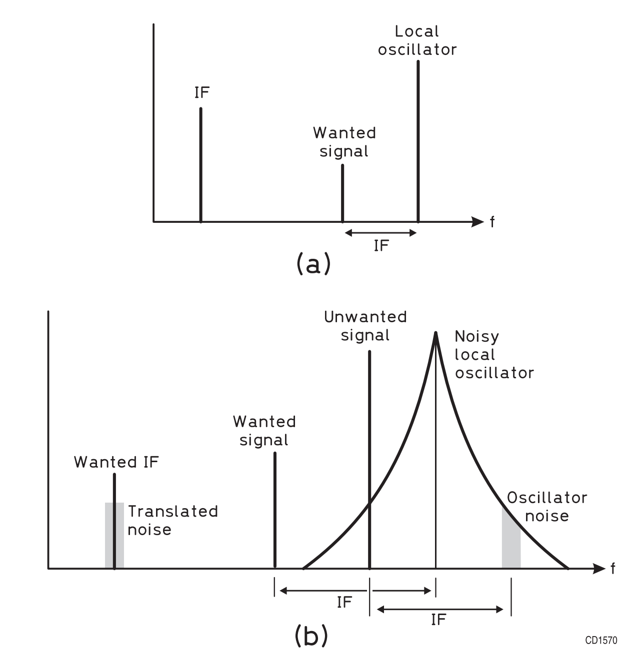


Fig 6.26: Superhet mixing. (a) Local oscillator separated by the IF from the wanted signal. (b) Local oscillator noise sideband separated by the IF from a strong unwanted signal

Fig 6.26 shows in simplified form the basic mechanism of reciprocal mixing. Because of the noise sidebands (or 'skirts') of the local oscillator, some part of the strong unwanted signal is translated into the receiver's IF pass-band and thus reduces the SNR of a weak wanted signal; further (not shown) a small fragment of the oscillator noise also spreads out to the IF, enters the IF channel and reduces the SNR. In practice the situation is even more complex since the very strong unwanted carrier will itself have noise sidebands which spread across the frequency of the wanted signal and degrade SNR no matter how pure the output of the local oscillator.

Fig 6.27 indicates the practical effect of reciprocal mixing on high-performance receivers; in the case of receiver 'A' (typical of many high-performance receivers) it is seen that the dynamic selectivity is degraded to the extent that the SNR of a very weak signal will be reduced by strong unwanted signals (1 to 10mV or more) up to 20kHz or more off-tune, even assuming that the front-end linearity is such that there is no cross-modulation, blocking or intermodulation. Reciprocal mixing thus tends to be the limiting factor affecting very weak station performance in real situations, although intermodulation or cross-modulation characteristics become the dominant factors with stronger signals.

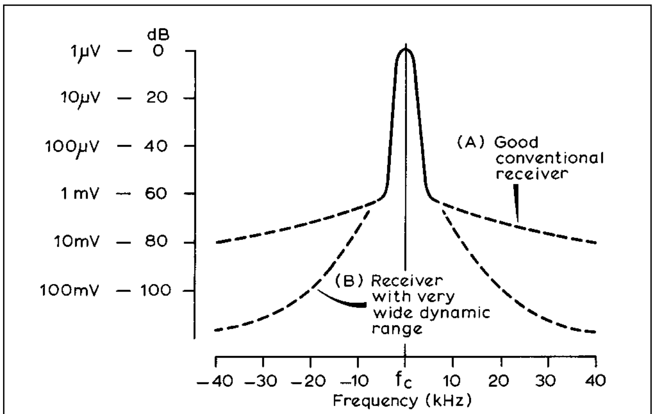


Fig 6.27: Reciprocal mixing due to oscillator noise can modify the overall selectivity curve of an otherwise very good receiver

The phase limited dynamic range of the receiver should therefore be approximately equal to the spurious-free dynamic range of the receiver: it is of no use having only one of them very good.

It is thus important in the highest-performance receivers to pay attention to achieving low noise-sidebands in oscillators, and this is one reason why the simpler forms of frequency synthesisers, which often have appreciable jitter and noise in the output, must be viewed with caution despite the high stability they achieve. The major noise mechanisms are:

(a) low frequency (1/f) noise which predominates close to carrier;
(b) thermal noise which defines the noise floor at frequency offsets greater than f/Q.

D B Leeson (ex-W6QHS, now W6NL) provided a simple model for oscillator noise [6]. Oscillator noise can be predicted as follows.

The basic oscillator noise floor is given by:

$$-174 + 3 + f \text{ dBm}$$

where f is the noise figure of the active device under self-oscillating limiting conditions. This is much higher than as a low-noise amplifier, and figures of 10dB for a transistor capable of 3dB as an amplifier are not uncommon.

From an offset from the carrier of f_o / Q (where f_o is the oscillator frequency, and Q is the working Q of the tank circuit), the noise rises at 6dB/octave (20dB/decade), until the point is reached where the 1/f noise starts to dominate. At offsets below this, the noise rises at 9dB/octave.

This 1/f knee is dependent upon the technology used: silicon FETs have very low 1/f noise knees: high-f_T bipolar transistors can have noise knees as high as 1kHz, while GaAs devices are measured in terms of megahertz.

A practical case is that of a transistor oscillator at 70MHz, running at 0dBm, with a 10dB noise figure, and a tank circuit Q of 70. From the above equations, the noise in a 1Hz bandwidth would -174 + 3 + 10 or -161dBm at 1MHz offset. At 100kHz, it would be -140dBm/Hz, at 10kHz, -120dBm/Hz, and at 1kHz -100dBm/Hz. The total amount of noise in the receiver IF bandwidth would be increased by the bandwidth: a 3kHz IF bandwidth receives 34dB more noise than a 1Hz bandwidth, so from a strong signal 10kHz away, 120 - 34dB or -86dB noise relative to wanted signal would be received. A filter with a 90dB stopband would be compromised by this level of noise.

It can be seen that the selectivity of a receiver can be determined not by the IF filters, but by the phase noise of the local oscillator.

Since the shot noise of an oscillator spreads across the IF channel, care is needed with low-noise mixers to limit the amount of this injection-source noise that enters the IF amplifier. Balanced and double-balanced mixers provide up to about 30dB rejection of oscillator noise.

Note that this noise only affects the sensitivity of the receiver; it is not related to reciprocal mixing. Another technique is to use a rejector trap (tuned circuit) resonant at the IF between oscillator and mixer.

Optimum oscillator performance calls for the use of a high unloaded tank-circuit Q. For switching-mode mixers and product detectors the optimum oscillator output waveform would be a square wave but this refinement is comparatively rare in practice.

Much more information on oscillators can be found in the first Building Blocks chapter.

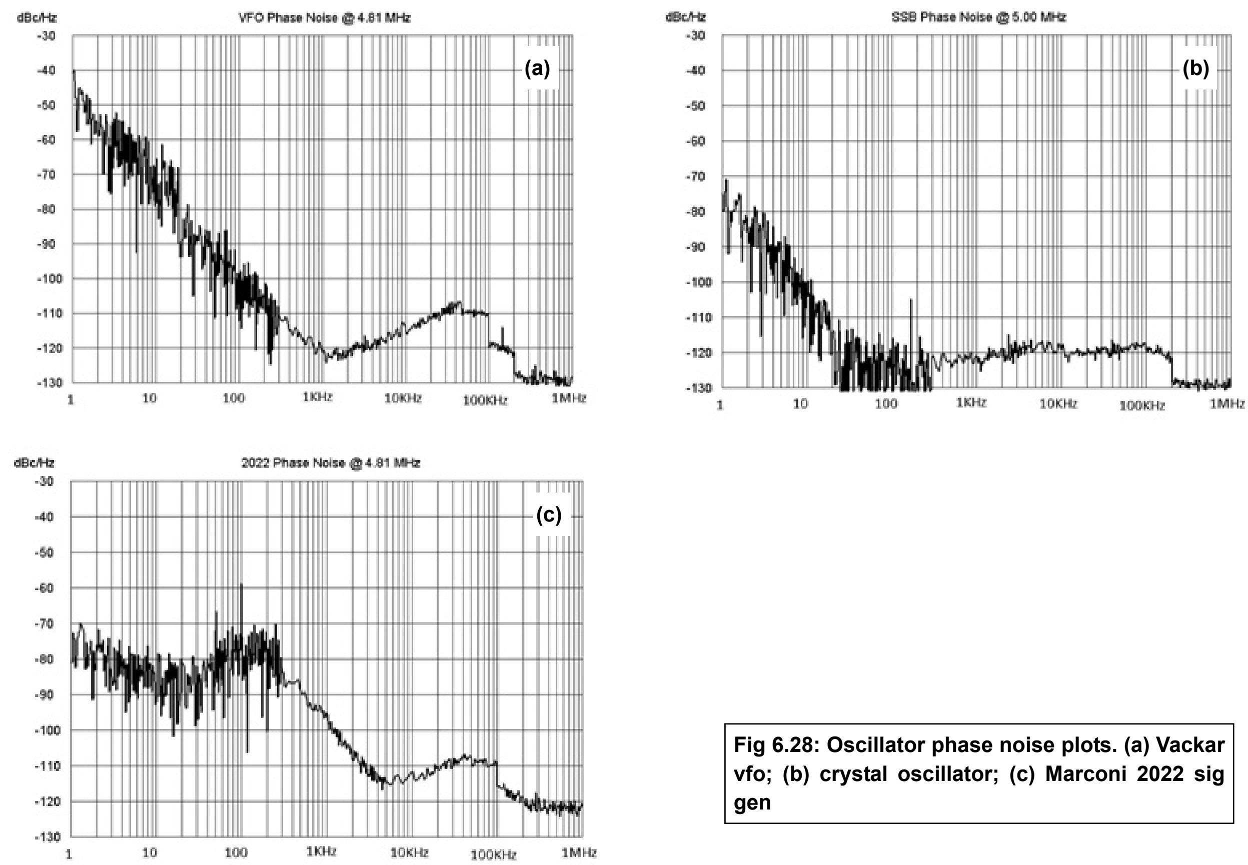


Fig 6.28: Oscillator phase noise plots. (a) Vackar vfo; (b) crystal oscillator; (c) Marconi 2022 sig gen

Fig 6.28 shows the measured phase noise of three types of oscillator, illustrating the wideband noise associated with each type. **(a)** shows the creditable performance of a well designed VFO as compared to the crystal oscillator in **(b)**. **(c)** is of a Marconi 2022 signal generator and illustrates the noise pedestal characteristic of frequency synthesisers.

Frequency Stability

The ability of a receiver to remain tuned to a wanted frequency without drift depends upon the electrical and mechanical stability of the internal oscillators. The primary cause of instability in an oscillator is a change of temperature, usually as the result of internally generated heat. With valves, even in the best designs, there will usually be steady frequency variation of any oscillator using inductors and capacitors during a period of perhaps 15 minutes or more after first switching on; transistor and FET oscillators reach thermal stability in a few seconds, provided that they are not then affected by other local sources of heat.

After undergoing a number of heat cycles, some components do not return precisely to their original values, making it difficult to maintain accurate calibration over a long period. Some receivers include a crystal-controlled oscillator of high stability providing marker signals (for example every 100kHz) so that receiver calibration can be checked and brought into adjustment.

Receiver drift can be specified in terms of maximum drift in Hertz over a specified time, usually quoting a separate figure to cover the warming-up period. For example a high-stability receiver might specify drift as "not worse than 200Hz in any five-hour period at constant ambient temperature and constant mains voltage after one hour warm-up".

Mechanical instability, which may appear as a shift of frequency when the receiver is subjected to mechanical shock or vibration, cannot easily be defined in the form of a performance specification. Sturdy construction on a mechanically rigid chassis can help. The need for high stability for SSB reception has led to much greater use of crystal-controlled oscillators and various forms of frequency synthesis.

Spurious Responses

A most important test of any receiver is the extent to which it receives signals when it is tuned to frequencies on which they are not really present, so adding to interference problems and misleading the operator. Every known type of receiver suffers from various forms of 'phantom' signals but some are much worse than others. Unfortunately, the mixing process is inherently prone to the generation of unwanted 'products' and receiver design is concerned with minimising their effect rather than their complete elimination. Spurii may take the form of:

(a) external signals heard on frequencies other than their true frequency;
(b) external signals which cannot be tuned out but are heard regardless of the setting of the tuning knob;
(c) carriers heard within the tuning range of the receiver but stemming not from external signals but from the receiver's own oscillators ('birdies').

Cases (a) and (b) are external spurious responses, while case (c) is an internal spurious.

In any superhet receiver tuneable signals may be created in the set whenever the interfering station or one of its harmonics (often produced within the receiver) differs from the intermediate frequency by a frequency equal to the local oscillator or one of its harmonics. This is reflected in the general expression:

$$mf_u \pm nf_0 = f_i$$

where m, n are any integers, including 0, f_u is the frequency of the unwanted signal, f_0 is the frequency of the local oscillator, and f_i is the intermediate frequency.

An important case occurs when m and n are 1, giving $f_u \pm f_0 = f_i$. This implies that f_u will either be on the frequency to which the set is correctly tuned (f_s) or differs from it by twice the IF ($2 \times f_i$), producing the so-called 'image' frequency. This is either $f_u = f_s + 2f_i$ (for cases where the local oscillator is higher in frequency than the wanted signal), or $f_u = f_s - 2f_i$ (where the local oscillator is lower in frequency than the wanted signal).

For an example, take a receiver tuned to about 14,200kHz with an IF of 470kHz and the oscillator high (ie about 14,670kHz). Such a receiver may, because of 'image', receive a station operating on 14,200 + (2 x 470) = 15,140kHz. Since 15,140kHz is within the 19m broadcast band, there is thus every likelihood that as the set is tuned around 14,200kHz strong broadcast signals will be received.

To reduce such undesirable effects, pre-mixer selectivity must be provided in the form of more RF tuned circuits or RF band-pass filters, or by increasing the Q of such circuits, or alternatively by increasing the frequency difference between the wanted and unwanted image signals. This frequency separation can be increased by increasing the factor $2f_i$, in other words by raising the intermediate frequency, and so allowing the broadly tuned circuits at signal frequency to have more effect in reducing signals on the unwanted image frequency before they reach the mixer.

It should also be noted from the general formula $mf_u \pm nf_0$ that 'image' is only one (though usually the most important) of many possible frequency combinations that can cause unwanted signals to appear at the intermediate frequency, even on a single-conversion receiver. The problem is greatly increased when more than one frequency conversion is employed.

Even with good pre-mixer selectivity it is still possible for a number of strong signals to reach the mixer, drive this or an RF stage into non-linearity and then produce a series of intermodulation products as spurious signals within an amateur band (see Fig 6.21).

The harmonics of the HF oscillator(s) may beat against incoming signals and produce output in the IF pass-band; strong signals may generate harmonics in the receiver stages and these can be received as spurii. Most such forms of spurii can be reduced by increasing pre-mixer selectivity to decrease the number of strong signals reaching the mixer, by increasing the linearity of the early stages of the receiver, or by reducing the amplitude of signals within these stages by the use of an antenna attenuator.

The case of internal spurious responses is similar. Here signals at $mf_1 + nf_2$ equalling the first IF or, more rarely, the tune or image frequency or the second IF can cause problems. Where a third or further IF is used, the oscillators there have also to be added to the equations. Usually, careful screening and filtering and particular attention to earth loops are required to minimise these problems in multi-conversion general-coverage receivers. In receivers for the amateur bands, careful choice of IF is necessary to minimise problems, although not even then can such difficulties be entirely eliminated. A typical 'built in' problem of this type is the use of a 'backwards tuning' 5 - 5.5MHz VFO, producing a spur from its 4th harmonic on 21.2MHz.

Very strong signals on or near the IF may break directly into the IF amplifier and then appear as untunable interference. This form of interference (although it then becomes tuneable) is particularly serious with the variable IF type of multiple-conversion receiver since there is almost certain to be a number of very strong signals operating over the segment of the HF spectrum chosen to provide the variable IF. Direct breakthrough may occur if the screening within the receiver is insufficient or if signals can leak in through the early stages due to lack of pre-mixer selectivity. For single-conversion and double-conversion with fixed first IF, it is common practice to include a resonant 'trap' (tuned to the IF) to reject incoming signals on this frequency. The multiple-conversion superhet contains more internal oscillators, and harmonics of the second and third (and occasionally the BFO) can be troublesome. For amateur-bands-only receivers every effort should be made to choose intermediate and oscillator frequencies that avoid as far as possible the effects of oscillator harmonics (birdies).

Ideally one would like an IF rejection (compared with the wanted signal) of almost 120dB but most amateurs would be well satisfied with 80-100dB of protection; in practice many receivers with variable first IF do not provide more than about 40-60dB protection.

Because of the great difficulty in eliminating spurious responses in double- and triple-conversion receivers, the modern designer tends to think more in terms of single conversion with high IF (eg 9MHz). Potentially the direct-conversion receiver is even more attractive, though it needs to be fairly complex to eliminate the audio-image response. It must also have sufficient linearity or pre-detector selectivity to reduce any envelope detection of very strong signals which may otherwise break through into the audio channel, regardless of the setting of the heterodyne oscillator. The direct-conversion receiver can also suffer from spurii resulting from harmonics of the signal or oscillator and this needs to be reduced by RF selectivity. A difficulty that can arise is that close in phase noise on the local oscillator being transferred by leakage to the mixer input, and then treated as an input signal, being demodulated, resulting in the noise appearing at AF.

Specifying the Receiver Performance

Few international standards specify the performance of amateur radio receivers. Standards makers are only concerned with the interference that such equipment might cause to other co-located equipment and there is no interest in how well the receiver performs its primary function. Therefore manufacturers are largely free to set their own specifications, which include the cost. We also have the customers of such equipment who at some point have to make a choice, probably based on cost, recommendations, reviews and advertising. Finally there is the designer/constructor who strives for high performance at any cost or the best performance for a minimum cost or technical novelty or any number of personal goals.

Often the performance is a function of the architecture or devices chosen at the beginning, with the final specification emerging at the end of the project, and this approach is entirely in the spirit of amateur radio. However, when considering major projects there is the other way where the specification comes first and the design evolves to meet it. The challenge for the designer/constructor is to set a specification that is meaningful, affordable and achievable.

The basic job of a radio is to receive and demodulate a signal at a prescribed frequency, so the top level performance parameters are sensitivity and frequency. All the other receiver performance parameters are related to how well it doesn't do things, usually referred to as rejection ratios eg adjacent channel rejection and image rejection. Consequently, with the exception of sensitivity and frequency, a list of such parameter measurements will each be several tens of decibels. Table 2 shows a list of receiver parameters and their definitions.

If you measured the performance of a cheap short wave radio, a lot of its parameters would be at the 40dB level, whilst measuring a professional receiver would probably produce results in the 100dB range. The cheap short wave radio is unlikely to have any feature that achieves 100dB similarly the professional receiver would not be spoilt with a 40dB feature. So to an approximation you can define the performance of a receiver by the number of decibels delivered by most of its parameters, and the cheap short wave receiver is a 40dB radio whilst the professional receiver is a 100dB radio.

For the designer/constructor, deciding to build a 70dB radio or a 90dB radio is all that is necessary because the rest i.e. performance, specifications, complexity and cost are inherent in that decision. Similarly, commercial receivers could be classified in the same way so as to give the prospective buyer a better way of choosing.

PARAMETER	PROPOSED DEFINITION
Reference sensitivity	The input level (dBm) required to produce a 10dB S/N output
Reference output	100mW into 8 ohms
AGC range	The change in input (dB) above the reference sensitivity that produces a change of 3dB in the reference output
Spurious free dynamic range	Derived from the IIP3 minus the minimum discernible input signal range
Blocking	The number of decibels above the reference sensitivity an unwanted off channel (by 20KHz) CW signal must be to cause a 3dB drop in a wanted reference output
Cross-modulation	The number of decibels above the reference sensitivity that an unwanted off channel signal* must be to cause a 3dB change in the wanted reference output. [* the unwanted signal is 20KHz off-channel and 50% amplitude modulated]
Image rejection	The input level (dB) above the reference sensitivity level that an input at the image frequency(s) must be to give the reference output level
IF rejection	The input level (dB) above the reference sensitivity level that an input at the IF must be to give the reference output level
Local oscillator leakage	Conducted power measurement at the LO frequency(s) coming out of the antenna socket
Adjacent channel selectivity	The input level (dB) above the reference sensitivity that an adjacent channel signal must be to produce the reference output
Internally generated spurious signals	The input level (dBm) for the equivalent to the highest internally generated spurious signal

Table 6.2: Receiver parameter definitions

Choice of IF

Choice of the intermediate frequency or frequencies is a most important consideration in the design of any superhet receiver. The lower the frequency, the easier it is to obtain high gain and good selectivity and also to avoid unwanted leakage of signals round the selective filter. On the other hand, the higher the IF, the greater will be the frequency difference between the wanted signal and the 'image' response, so making it simpler to obtain good protection against image reception of unwanted signals and also reducing the 'pulling' of the local oscillator frequency. These considerations are basically opposed, and the IF of a single-conversion receiver is thus a matter of compromise; however, in recent years it has become easier to obtain good selectivity with higher-frequency band-pass crystal filters and it is no longer any problem to obtain high gain at high frequencies. The very early superhet receivers used an IF of about 100kHz; then for many years 455-470kHz was the usual choice - many modern designs use between about 3 and 9MHz, and SSB IF filters are now available to 40MHz. A good rule of thumb is that the IF should be no less than 5%, and preferably 10% of the signal frequency. Where the IF is higher than the signal frequency the action of the mixer is to raise the frequency of the incoming signal, and this process is now often termed up-conversion (a term formerly reserved for a special form of parametric mixer). Up-conversion, in conjunction with a low-pass filter at the input, is an effective means of reducing IF breakthrough as well as image response.

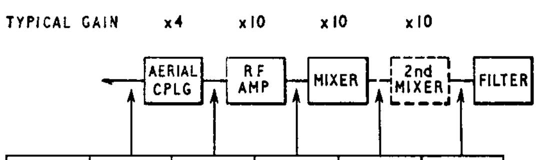
A superhet receiver, whether single- or multi-conversion, must have its first IF outside its tuning range. For general-coverage HF receivers tuning between, say, 1.5 and 30MHz, this limits the choice to below 1.5MHz or above 30MHz. To reduce image response without having to increase pre-mixer RF selectivity (which can involve costly gang-tuned circuits) professional designers are increasingly using a first IF well above 30MHz. This trend is being encouraged by the availability of VHF crystal filters suitable for use either directly as an SSB filter or more often with relaxed specification as a roofing filter.

The use of a very high first IF, however, tends to make the design of the local oscillator more critical (unless the Wadley triple-conversion drift-free technique is used - when the intermodulation problems are more difficult because of the number of stages prior to the selectivity determining filters). For amateur-bands-only receivers the range of choice for the first IF is much wider, and 3.395MHz and 9MHz are typical.

A number of receivers have adopted 9MHz IF with 5.0-5.5MHz local oscillator: this enables 4MHz-3.5MHz and 14.0-14.5MHz to be received without any band switching in the VFO; other bands are received using crystal-controlled converters with outputs at 3.5 or 14MHz. With frequency synthesisers, up-conversion is commonly found, with the first IF between 40-70MHz.

Gain Distribution

In many receivers of conventional design, it has been the practice to distribute the gain throughout the receiver in such a manner as to optimise signal-to-noise ratio and to minimise spurious responses. So long as relatively noisy mixer stages were used it was essential to amplify the signal considerably before it reached the mixer. This means that any strong unwanted signals, even when many kilohertz from the wanted signal, pass through the early unselective amplifiers and are built up to levels where they cause cross-modulation within the mixer: see **Fig 6.29**. Today it is recognised that it is more satisfactory if pre-mixer gain can be kept low to prevent this happening. Older multi-grid valve mixers had an equivalent noise resistance as high as 200,000Ω (representing some 4-5mV of noise referred to the grid). Later types such as the ECH81 and 6BA7 reduced this to an ENR of



WANTED SIGNAL	1µV	4µV	40µV	400µV	4mV
UNWANTED SIGNAL	1mV	4mV	40mV	0·4V	4V
UNWANTED SIGNAL	10mV	40mV	0·4V	4V	40V
UNWANTED SIGNAL	50mV	200mV	2V	20V	200V

Fig 6.29: This diagram shows how unwanted signals are built up in high-gain front-ends to levels at which cross-modulation, blocking and intermodulation are virtually bound to occur

about 60,000Ω (about 2.25mV of noise) while the ENR of pentode and triode mixers was lower still (although these may not be as satisfactory for mixers in other respects).

The noise contribution of semiconductor mixers is also low - for example an FET mixer may have a noise factor as low as 3dB - so that generally the designer need no longer worry unduly about the requirement for pre-mixer amplification to overcome noise problems. Nevertheless a signal frequency stage may still be useful in helping to overcome image reception by providing a convenient and efficient method of coupling together signal-frequency tuned circuits, and when correctly controlled by AGC it becomes an automatic large-signal attenuator. Pre-mixer selectivity limits the number of strong signals reaching the mixer. The use of double-balanced mixers is generally advisable to reduce the spurious response possibilities.

Fig 6.30 shows a typical gain distribution as found in a modern design in which the signal applied to the first mixer is much lower than was the case with older (valved) receivers.

The significant conversion losses of modern diode and FET ring mixers (6-10dB), the losses of input band-pass filtering and the need for correct impedance termination means that in the highest-performance receivers it is desirable to include low-gain, low-noise, high dynamic range pre- and post-mixer amplifiers and a diplexer ahead of the (main or roofing) crystal filter to achieve constant input impedance over a broad band of frequencies.

If the diplexer is a simple resistive network this will introduce a further loss of some 6dB. Stage-by-stage gain, noise figure, third-order intercept, 1dB compression point and 1dB desensitisation point performance of the N6NWP high-dynamic-range MF/HF front-end of a single-conversion (9MHz IF) receiver [7](QST February 1993, and see later) is shown in **Fig 6.31**. More complex diplexers may be used to divert the local oscillator feedthrough signal from entering the IF strip.

For double- and multiple-conversion receivers, the gain distribution and performance characteristics of all the stages preceding the selective filter need to be considered. In general the power loss in any band-pass filter will decrease as the bandwidth increases: for example a 75MHz, 25kHz-bandwidth filter might have a power loss of 1dB whereas a 75MHz, 7kHz-bandwidth filter might have a power loss of 3.5dB.

FREQUENCY STABILITY OF RECEIVERS

The resolution of SSB speech and the reception of a CW signal requires that a receiver can be tuned to, and remain within, about 25-30Hz of the frequency of the incoming signal. At 29MHz this represents a tolerance of only about one part in a million. For amateur operation the main requirement is that this degree of stability should be maintained over periods of up to about 30min. Long-term stability is less important for amateurs than short-term stability; it will also be most convenient if a receiver reaches this degree of stability within a fairly short time of switching on.

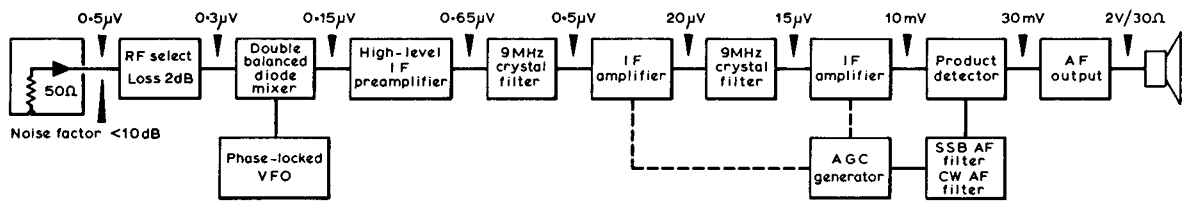
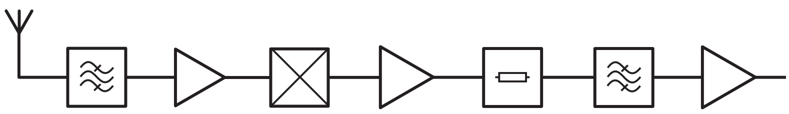


Fig 6.30: Gain distribution in a high-performance semiconductor single-conversion receiver built by G3URX using seven integrated circuits, 27 transistors and 14 diodes. 1µV signals can be received 5kHz off-tune from a 60mV signal and the limiting factor for weak signals is the noise sidebands of the local oscillator, although the phase-locked VFO gives lower noise and spurious responses than the more usual pre-mixer VFO system



	Input band-pass filter	Preamp	Mixer	Post-mixer amp	−6dB network	9MHz xtal filter	IF amp
Gain (A_V) (dB)	−0·3	8	−7	8	−6	−2	–
Noise figure (NF) (dB)	0·3	2	7	2·5	6	2	3
Third-order input intercept (IP_3) (dBm)	+33	+32·7	+40·7	+34·5	–	–	–
1dB input compression point (P_C) (dBm)	+15·3	+15	+23	+18	–	–	–
1dB input desens point (P_D) (dBm)	+14·3	+14	+22	–	–	–	–

Fig 6.31: Stage by stage gain and characteristics of the N6NWP front-end of single-conversion 14MHz HF receiver with 9MHz IF

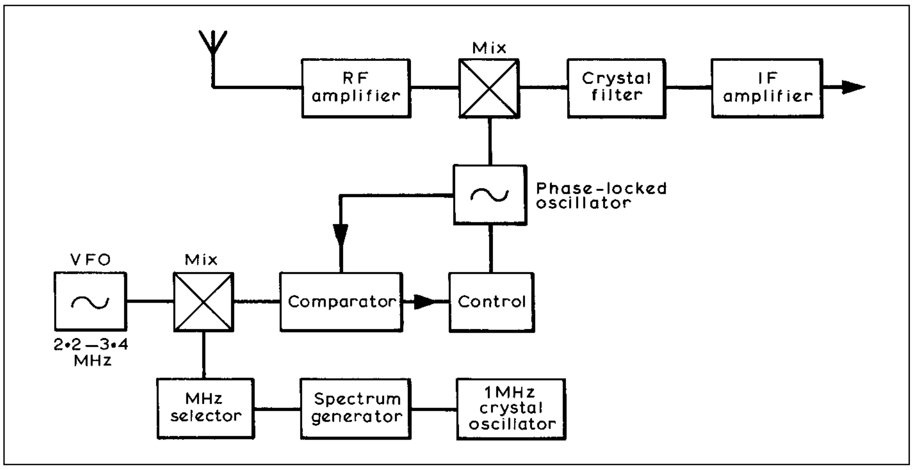


Fig 6.32: Practical frequency synthesis using fixed-range VFO with megahertz signals derived from a single 1MHz crystal

It is extremely difficult to achieve or even approach this order of stability with a free-running, band-switched variably tuned oscillator working on the fundamental injection frequency, although with care a well-designed FET oscillator can come fairly close. This has led (in the same way as for transmitting VFO units) to various frequency-synthesis techniques in which the stability of a free-running oscillator is enhanced by the use of crystals. The following are among the techniques used:

Multi-conversion Receiver with Crystal-controlled First Oscillator & Variable First IF

This very popular technique has a tuneable receiver section covering only one fixed frequency band; for example 5.0-5.5MHz. The oscillator can be carefully designed and temperature compensated over one band without the problems arising from the uncertain action of wavechange switches, and can be separated from the mixer stage by means of an isolating or buffer stage. For the front-end section a separate crystal is needed for each tuning range (the 28MHz band may require four or more crystals to provide full coverage of 28.0-29.7MHz).

Partial Synthesis

The arrangement of (1) becomes increasingly costly to implement as the frequency coverage of the variable IF section is reduced below about 500kHz, or is required to provide general coverage throughout the HF band. Beyond a certain number of crystals it becomes more economical (and offers potentially higher stability) if the separate crystals are replaced by a single high-stability crystal (eg 1MHz) from which the various band-setting frequencies are derived (**Fig 6.32**). This may be done, for example, by digital techniques or by providing a spectrum of harmonics to one of which a free-running oscillator is phase-locked. Note that with this system the tuning within any band still depends on the VFO and for this reason is termed partial synthesis.

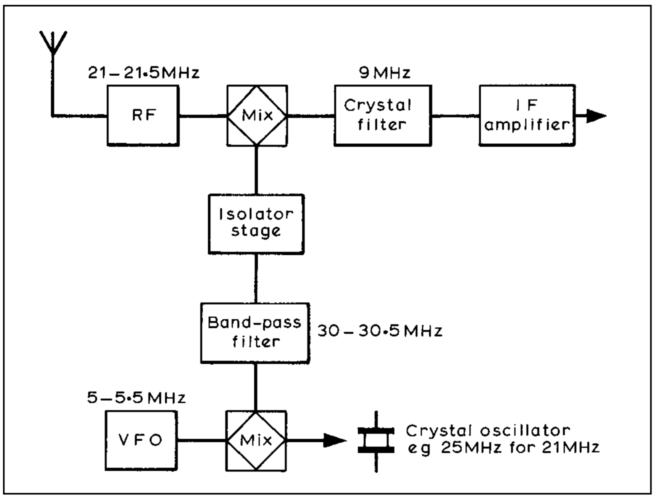


Fig 6.33: Pre-mixer heterodyne VFO system provides constant tuning rate with single-conversion receiver. Requires a number of crystals and care must be taken to reduce spurious oscillator products reaching the main mixer

Single-conversion Receiver with Heterodyne (Pre-mixer) VFO

In this system of partial synthesis, the receiver may be a single-conversion superhet (or dual-conversion with fixed IFs) (**Fig 6.33**). The variable HF injection frequency is obtained using a heterodyne-type VFO, in which the output of a crystal-controlled oscillator is mixed with that of a single-range VFO, and the output is then filtered and used as the injection frequency. The overall stability will be much the same as for (1) but the system allows the selective filter to be placed immediately after the first mixer stage. However, to reduce spurious responses the unwanted mixer products of the heterodyne-VFO must be reduced to a very low level and not reach the mixer. As with the tuneable IF system, this arrangement results in equal tuning rates on all bands. The system can be extended by replacing the series of separate crystals with a single crystal plus phase-locking arrangement as in the Partial Synthesis case above.

Fixed IF Receiver with Partial Frequency Synthesis

Fig 6.34 outlines an ingenious frequency synthesiser (due to Plessey) incorporating an interpolating LC oscillator and suitable for use with single- or multiple-conversion receivers having fixed intermediate frequencies. The output of the VFO is passed through a variable-ratio divider and then added to the reference frequency by means of a mixer. The sum of the two frequencies

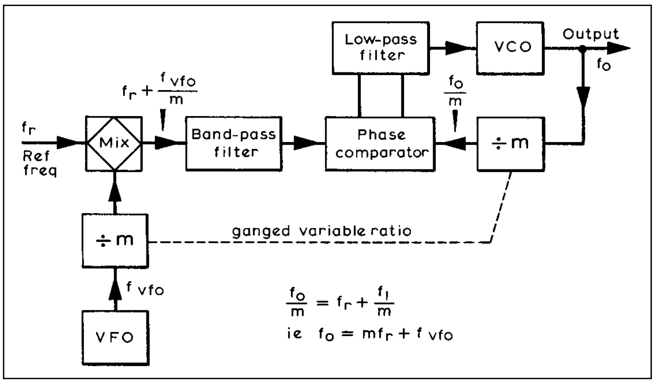


Fig 6.34: A digital form of partial synthesis developed by Plessey and suitable for use in single-conversion receivers

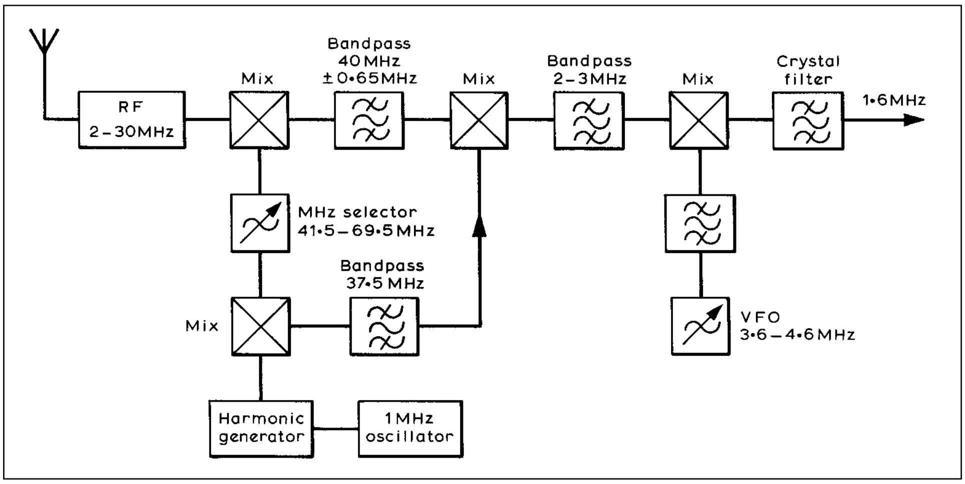


Fig 6.35: The Wadley drift-cancelling loop system as used on some early Racal HF receivers but requiring a considerable number of mixing processes and effective VHF band-pass filters

applied to the mixer is selected by means of a band-pass filter and provides one input to a phase comparator; the other input to this phase comparator is obtained from the output of the voltage-controlled oscillator after it has also been divided in the same ratio as the interpolating frequency. The phase comparator can then be used to phase lock the VCO to the frequency $mf_{ref} + f_{VFO}$ where m represents the variable-ratio division. If for example f_{ref} is 1MHz, f_{VFO} covers a tuning range of 1-2MHz and the variable ratio dividers are set to 16, then the VCO output can be controlled over 17,000-18,000kHz; if the ratio divider is changed to 20 then the tuning range becomes 21,000-22,000kHz and so on. The use of two relatively simple variable ratio divider chains thus makes it possible to provide output over the full HF range, with the VFO at a low frequency (eg 1-2MHz).

VXO Local Oscillator

For reception over only small segments of a band or bands, a variable-crystal oscillator (VXO) can be used to provide high stability for mobile or portable receivers. As explained in the chapter on oscillators, the frequency of a crystal can be 'pulled' over a small percentage of its nominal frequency without significant loss of stability. The system is attractive for small transceivers. Oscillators based on ceramic resonators can be 'pulled' over significantly greater frequency ranges.

Drift-cancelling Wadley Loop

A stable form of front-end for use with variable IF-type receivers is the multiple-conversion Wadley loop which was pioneered in the Racal RA17 receiver (**Fig 6.35**). By means of an ingenious triple-mixing arrangement a variable oscillator tuning 40.5 to 69.5MHz and a 1MHz crystal oscillator provides continuous tuning over the range 0.5 to 30MHz as a series of 1MHz segments. Any drift of the variable VHF oscillator is automatically corrected. Although the system has been used successfully in home-constructed receivers, it is essential to use a good VHF band-pass filter (eg 40MHz ± 0.65MHz) and extremely good screening if spurious responses are to be minimised.

PLL Frequency Synthesisers and Direct Digital Synthesis (DDS)

These systems produce an output waveform which is locked precisely to a crystal reference frequency. The frequency stability is therefore equal to that of the crystal oscillator. Both these techniques are described in detail in the chapter on oscillators.

ACTIVE DEVICES FOR RECEIVERS

The radio amateur is today faced with a wide and sometimes puzzling choice of active devices around which to design a receiver: valves, bipolar transistors, field-effect transistors including single- and dual-gate MOSFETs and junction FETs; special diodes such as Schottky (hot-carrier) diodes, and an increasing number of integrated circuits, many designed specifically for receiver applications.

Each possesses advantages and disadvantages when applied to high-performance receivers, and most recent designs tend to draw freely from among these different devices.

Valves

In general, valves are bulky, require additional wiring and power supplies for heaters, generate heat, and are subject to ageing in the form of a gradual change of characteristics throughout their useful life. On the other hand they are not easily damaged by high-voltage transients; were manufactured in a wide variety of types for specific purposes to fairly close tolerances; and are capable of handling small signals with good linearity (and special types can cope well with large signals, though mixers are noisy). See the chapter on Semicondctors and Vaves for more.

Bipolar Transistors

These devices can provide very good noise performance with high gain, are simple to wire and need only low-voltage supplies, consuming very little power and so generating very little heat (except power types needed to form the audio output stage). On the other hand, they are low-impedance, current-operated devices, making the interstage matching more critical and tending to impose increased loading on the tuned circuits; they have feedback capacitances that may require neutralising; they are sensitive to heat, changing characteristics with changing temperature; and they can be damaged by large input voltages or transients. Their main drawback in the signal path of a receiver is the difficulty of achieving wide dynamic range and satisfactory AGC characteristics. On the other hand, bipolar transistors are suitable for most AF applications.

The bipolar transistors developed for CATV (wire distribution of TV) such as the BFW17, BFW17A, 2N5109 etc, used with a heatsink, can form excellent RF stages or mixers, especially where feedback is used to enhance linearity. Not all transistors are equal in linearity, and those developed for CATV show advantages.

Field-effect Transistors

These devices can offer significant advantages over bipolar devices for the low-level signal path. Their high input impedance makes accurate matching less important; their near square-law characteristics make them comparable with variable-mu valves in reducing susceptibility to cross-modulation; at the same time, second-order intermodulation and responses to signals at 'half the IF' removed necessitate careful consideration of front-end filtering.

They can readily be controlled by AGC systems, although attempts to obtain too much gain reduction can lead to severe distortion. The dual-gate form of device is particularly useful for small-signal applications, and forms an important device for modern receivers. They tend, however, to be limited in signal-handling capabilities. Special types of high-current field-effect transistors have been developed capable of providing extremely wide dynamic range (up to 140dB) in the front-ends of receivers. A problem with FET devices is the wide spread of characteristics between different devices bearing the same type number, and this may make individual adjustment of the bias levels of FET stages desirable. The good signal-handling capabilities of MOSFET mixers can be lost by incorrect signal or local oscillator levels.

Integrated Circuits

Special-purpose integrated circuits use large numbers of bipolar transistors in configurations designed to overcome many of the problems of circuits based on discrete devices. Because of the extremely high gain that can be achieved within a single IC they also offer the home-constructor simplification of design and construction. High-performance receivers can be designed around a few special-purpose linear ICs, or one or two consumer-type ICs may alternatively form the 'heart' of a useful communications receiver.

It should be recognised, however, that their RF signal-handling capabilities are less than can be achieved with special-purpose discrete devices and their temperature sensitivity and heat generation (due to the large number of active devices in close proximity) usually make them unsuitable for oscillator applications. They also have a spread of characteristics which may make it desirable to select devices from a batch for critical applications.

The electronics industry has tended towards digital techniques where possible, and generally to application specific integrated circuits (ASICs) for both analogue and digital applications. In general, such devices as operational amplifiers and some audio amplifiers will continue to be available, but many of the small-scale integration (SSI) devices that were the mainstay of the amateur throughout the 1970s and 1980s have now disappeared. Of the circuits manufactured for radio applications, many are aimed at markets such as cellular radio, and have become so dedicated that their use (and availability in small quantities) for amateur radio is doubtful.

Integrated-circuit precautions

As with all semiconductor devices it is necessary to take precautions with integrated circuits, although if handled correctly high reliability may be expected.

Recommended precautions include:

- Do not use excessive soldering heat and ensure that the tip is at earth potential
- Take precautions against static discharge - eg don't walk across a synthetic fibre carpet on a cold day and then touch a device. It is advisable if handling CMOS devices to use a static bleed wriststrap, connected to earth.
- Check and recheck all connections several times before applying any voltages.
- Keep integrated circuits away from strong RF fields.
- Keep supply voltages within ±10% of those specified for the device from well-smoothed supplies.

Integrated circuit amplifiers can provide very high gains (eg up to 80dB or so) within a single device having input and output leads separated by only a small distance: this means that careful layout is needed to avoid instability, and some devices may require the use of a shield between input and output circuits. Some devices have earth leads arranged so that a shield can be connected across the underside of the device.

Earth returns are important in high-gain devices: some have input and output earth returns brought out separately in order to minimise unwanted coupling due to common earth return impedances, but this is not true of all devices.

Normally IC amplifiers are not intended to require neutralisation to achieve stability; unwanted oscillation can usually be traced to unsatisfactory layout or circuit arrangements. VHF parasitics may generally be eliminated by fitting a 10 ohm resistor in series with either the input or output lead, close to the IC.

With high-gain amplifiers, particular importance attaches to the decoupling of the voltage feeds. At the low voltages involved, values of series decoupling resistors must generally be kept low so that the inclusion of low-impedance bypass capacitors is usually essential. Since high-Q RF chokes may be a cause of RF oscillation it may be advisable to thread ferrite beads over one lead of any RF choke to reduce the Q.

As with bipolar transistors, IC devices (if based on bipolar transistors) have relatively low input and output impedances so that correct matching is necessary between stages. The use of a FET source follower stage may be a useful alternative to step-down transformers for matching.

Maximum and minimum operating temperatures should be observed. Many linear devices are available at significantly lower cost in limited temperature ranges which are usually more than adequate for operation under normal domestic conditions.

Because of the relatively high temperature sensitivity and noise (especially LF noise) of bipolar-type integrated circuits, they are not generally suitable as free-running oscillators in high-performance receivers.

For the very highest grade receivers, discrete components and devices are still required in the front-end since currently available integrated circuits do not have comparable dynamic range. The IC makes possible extremely compact receivers; in practice miniaturisation is now limited - at least for general-purpose receivers - by the need to provide easy-to-use controls for the non-miniaturised operator.

SELECTIVE FILTERS

The selectivity characteristics of any receiver are determined by filters: these filters may be at signal frequency (as in a straight receiver); intermediate frequency; or audio frequency (as in a direct-conversion receiver). Filters at signal frequency or IF are usually of band-pass characteristics; those at AF may be either band-pass or low-pass. With a very high first IF (112-150MHz) low-pass filters may be used at RF.

A number of different types of filters are in common use: LC (inductor-capacitor) filters as in a conventional IF transformer or tuned circuit; crystal filters; mechanical filters; ceramic filters; RC (resistor-capacitor) active filters (usually only at AF but feasible also at IF).

Roofing Filters

If the main selective filter is placed early in the signal path (eg immediately following a diode mixer or low-gain, low-noise, post-mixer amplifier) where the signal voltage is very low, subsequent amplification (of the order of 100dB or more) will introduce considerable broad-band noise unless further narrow-band filtering is provided (which may be AF filtering).

Another answer is to use an initial roofing filter with the final, more-selective filter(s) further down the signal path.

Crystal Filters

The selectivity of a tuned circuit is governed by its frequency and by its Q (ratio of reactance to resistance). There are practical limits to the Q obtainable in coils and IF transformers. In 1929, Dr J Robinson, a British scientist, introduced the quartz crystal resonator into radio receivers. The advantages of such a device for communications receivers were appreciated by James Lamb of the American Radio Relay League and he made popular the IF crystal filter for amateur operators.

For this application a quartz crystal may be considered as a resonant circuit with a Q of from 10,000 to 100,000 compared with about 300 for a very-high-grade coil and capacitor tuned circuit.

From earlier chapters, it will be noted that the electrical equivalent of a crystal is not a simple series- or parallel-tuned circuit, but a combination of the two: it has (a) a fixed series resonant frequency (f_s) and (b) a parallel resonant frequency (f_p). The frequency fp is determined partly by the capacitance of the crystal holder and by any added parallel capacitance and can be varied over a small range.

The crystal offers low impedance to signals at its series resonant frequency; a very high impedance to signals at its parallel resonant frequency, and a moderately high impedance to signals on other frequencies, tending to decrease as the frequency increases due to the parallel capacitance.

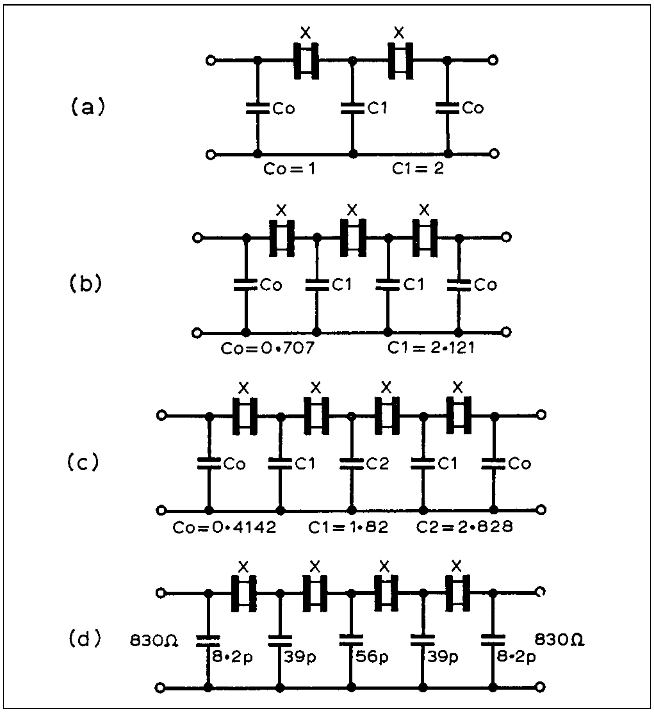


Fig 6.36: Crystal ladder filters, investigated by F6BQP, can provide effective SSB and CW band-pass filters. All crystals (X) are of the same resonant frequency and preferably between 8 and 10MHz for SSB units. To calculate values for the capacitors multiply the coefficients given above by $1/(2\pi fR)$ where f is frequency of crystal in hertz (MHz by 10^6), R is input and output termination impedance and 2π is roughly 6.8. (a) Two-crystal unit with relatively poor shape factor. (b) Three-crystal filter can give good results. (c) Four-crystal unit capable of excellent results. (d) Practical realisation of four-crystal unit using 8314kHz crystals, 10% preferred-value capacitors and termination impedance of 820Ω. Note that for crystals between 8 and 10MHz the termination impedance should be between about 800 and 1000Ω for SSB. At lower crystal frequencies use higher design impedances to obtain sufficient bandwidth. For CW filters use lower impedance and/or lower frequency crystals

Ladder crystal filters

Single or multiple crystal half-lattice filters were used in the valve era. The modern filter of choice is the ladder filter, which can provide excellent SSB filters at frequencies between about 4-11MHz. This form of crystal filter uses a number of crystals of the same (or nearly the same) frequency and so avoids the need for accurate crystal etching or selection. Further, provided it is correctly terminated, it does not require the use of transformers or inductors.

Plated crystals such as the HC6U or 10XJ types are more likely to form good SSB filters, although virtually any type of crystal may be used for CW filters.

A number of practical design approaches have been described by J Pochet, F6BQP, [8] and in a series of articles [9]. **Fig 6.36** outlines the F6BQP approach. By designing for lower termination impedances and/or lower frequency crystals excellent CW filters can be formed. A feature of the ladder design is the very high ultimate out-of-band rejection that can be achieved (75-95dB) in three- or four-section filters. For SSB filters at about 8MHz a suitable design impedance would be about 800Ω with a typical 'nose' band-pass of 2.0-2.1kHz. Intercept points of SSB and roofing filters can be over +50dBm with low insertion losses, and over +45dBm for narrow-band CW filters.

The ladder configuration is particularly attractive for home-built receivers since they can be based on readily available, low-cost 4.43MHz PAL colour-subcarrier crystals produced for use in domestic colour-TV receivers to P129 or P128 specifications. In NTSC countries, including North America and Japan, 3.58MHz TV crystals can be used, although this places the IF within an amateur band and would be unsuitable for single-conversion superhet receivers. Low-cost crystals at twice the PAL sub-carrier frequency (ie 8.86MHz) are also suitable.

Virtually any combination of crystals and capacitors produces a filter of some sort, but published equations or guidance

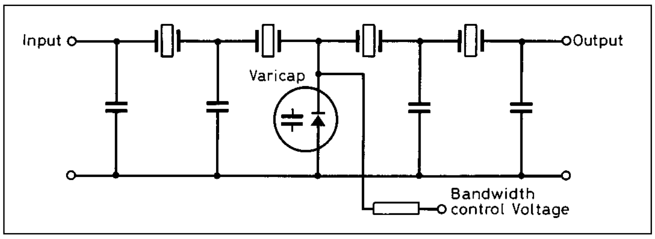


Fig 6.37: Simple technique for varying the bandwidth of an 8MHz crystal ladder filter from about 12.8kHz down to 1.1kHz

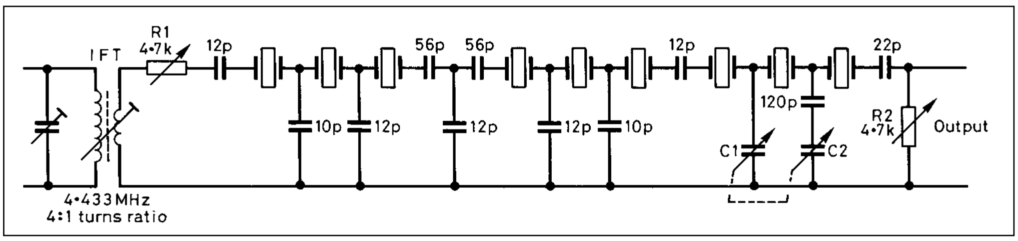


Fig 6.38: Variable-selectivity ladder filter using low-cost PAL colour-TV crystals for AM/SSB/CW/RTTY reception

should be followed to achieve optimum filter shapes and desired bandwidth. If this is done, ladder filters can readily be assembled from a handful of nominally identical crystals (ideally selected with some small offsets of up to about 50 or 100Hz) plus a few capacitors, yet providing SSB or CW filters with good ultimate rejection, plus reasonably low insertion loss and pass-band ripple.

Ladder filters have intercept points significantly above those of most economy-grade, lattice-type filters.

Figs 6.37 and 6.38 show typical ladder filter designs based on 4.43MHz and 8MHz crystals.

A valuable feature of the ladder configuration is that it lends itself to variable selectivity by changing the value of some, or preferably all, of the capacitors. The bandwidth can be varied over a restricted but useful range simply by making the middle capacitor variable, using a mechanically variable capacitor or electronic tuning diode (which may take the form of a 1W zener diode).

Fig 6.37 shows an 8MHz filter which has a 'nose' bandwidth that can be varied from about 2.8kHz down to 1.1kHz.

Fig 6.38 illustrates a 4.43MHz nine-crystal filter built by R Howgego, G4DTC, for AM/SSB/CW/RTTY reception; the 3dB points can be varied from 4.35kHz down to 600Hz. In development, he noted that the bandwidth is determined entirely by the 'vertical' capacitors. If, however, these are reduced below about 10pF, the bandwidth begins to narrow rather than widen. The maximum bandwidth that could be achieved was about 4.5kHz. This could be widened by placing resistors (1kΩ to 10kΩ) across the capacitors but this increases insertion loss. Terminating impedances affect the pass-band ripple, not the bandwidth.

The filter of Fig 6.38 gives continuously variable selectivity yet is relatively easy to construct. It is basically a six-pole roofing filter followed by a variable three-pole filter. It was based on low-cost Philips HC18-U type crystals and these were found to be all within a range of 80Hz.

Crystals in the large case style (eg HC6-U) tend to be about 200Hz lower. C1, C2 is a 60 + 142pF miniature tuning capacitor as found in many portable broadcast receivers. The integral trimmers are set for maximum bandwidth when capacitor plates are fully unmeshed. Set R1 for best compromise between minimum bandwidth and insertion loss, 1.2kΩ nominal.

The following specification should be achievable. C1, C2 plates unmeshed: 3dB points at 4437.25kHz and 4432.90kHz, bandwidth 4.35kHz. C1, C2 plates half-meshed: 3dB points at 4434.0kHz and 4432.90kHz, bandwidth 1.10kHz. C1, C2 plates meshed: 3dB points at 4433.5kHz and 4432.90kHz, bandwidth 600Hz. Insertion loss in pass-band: maximum (R1 2.5kΩ, R2 1.2kΩ) 10dB, minimum (R1 0kΩ , R2 1.2kΩ) 6dB. Pass-band ripple 1-3dB (dependent on R1). Ripple reduces with bandwidth. Stop-band attenuation better than 60dB. -20dB bandwidth typically 1kHz wider than the -3dB bandwidth.

The filter shown in Fig 6.39, designed by D Gordon-Smith, G3UUR, provides six different bandwidths suitable for both SSB and CW operation, switching the value of all capacitors, and ideally preceded by a roofing filter. In order to reduce the number of switched components, the terminating resistors remain constant and the ripple merely decreases with bandwidth. This also reduces the variation in insertion loss. The 2.4kHz position has a 1dB ripple Chebyshev response, and the 500Hz position represents a Butterworth response. A 5:1 bandwidth change is possible if the maximum tolerable ripple is 1dB. This range is fixed by design constraints: the ratio of 1dB to 0dB ripple response terminating resistance is approximately 5:1 for the same bandwidth, and therefore the same terminating resistance satisfies bandwidths that have a ratio of about 5:1. The main disadvantage is that the pass-band moves low in frequency as it is narrowed. This could be compensated for by moving the carrier crystals down in sympathy with the filter centre frequency; a total shift of 1kHz or less would probably be adequate, and could be corrected at the VFO by the RIT shift control.

A different approach to ladder filters is to put the crystals in shunt with the signal rather than in series. The filter then has its steeper slope on the low-frequency side instead of the high-frequency side, and this may be preferable for narrow-bandwidth filters (eg CW filters), particularly when using relatively low-Q plated crystals in HC-18 holders. Fig 6.40 shows a shunt-type crystal filter designed by John Pivnichy, N2DCH. The filter response is shown in Fig 6.41.

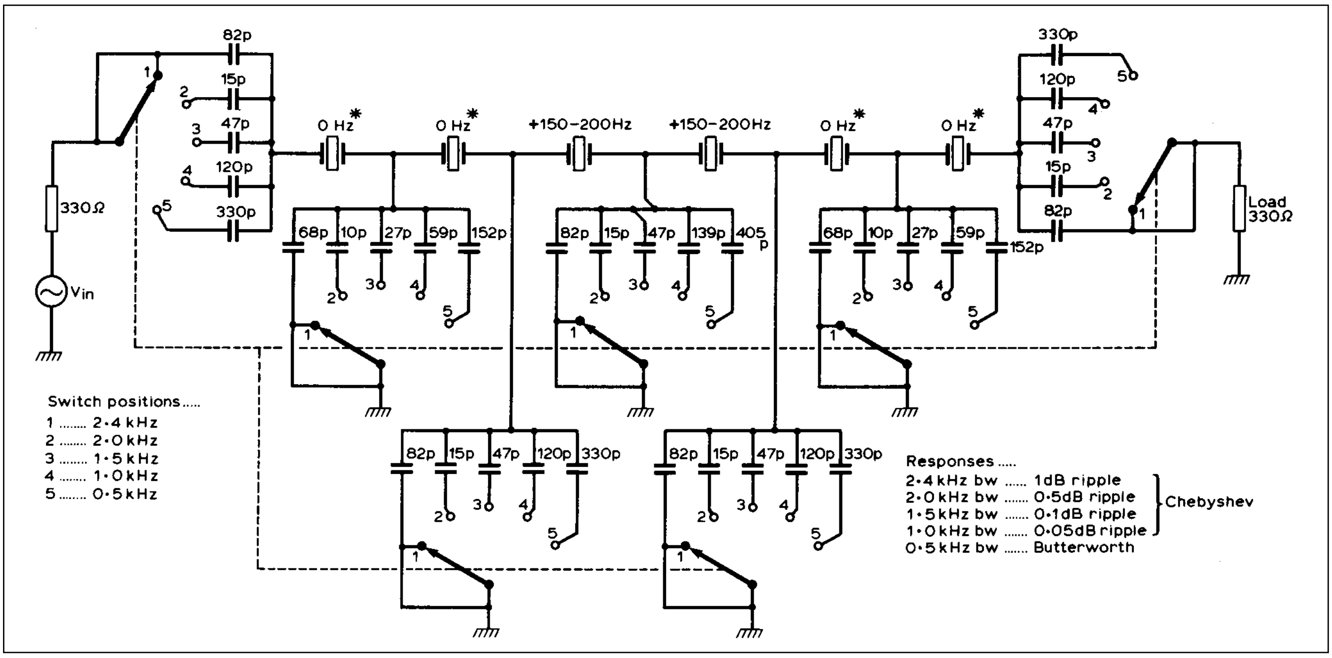


Fig 6.39: G3UUR's design for a switched variable-bandwidth ladder filter using colour-TV crystals. Note that crystals shown as 0Hz offset can be in practice ±0.5Hz without having too detrimental an effect on the pass-band ripple

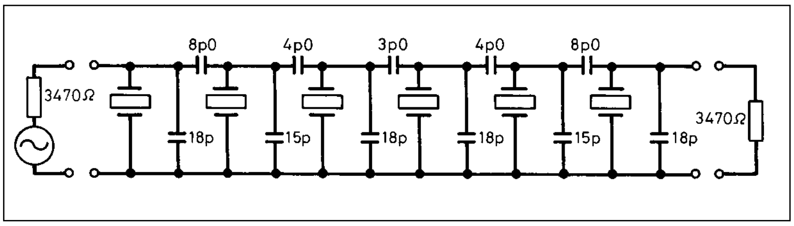


Fig 6.40: Shunt-type crystal ladder CW filter using six 3.58MHz NTSC crystals designed for 3470Ω terminations. Crystals should usually be matched to within 100Hz. (TV crystals are often only specified as within 300Hz but are usually within 200Hz)

Ceramic Filters

Piezoelectric effects are not confined to quartz crystals; in recent years increasing use has been made of certain ceramics, such as lead zirconate titanate (PZT). Small discs of PZT, which resonate in the radial dimension, can form economical selective filters in much the same way as quartz, though with considerably lower Q.

Ceramic IF transfilters are a convenient means of providing the low impedances needed for bipolar transistor circuits. The simplest ceramic filters use just one resonator but numbers of resonators can be coupled together to form filters of required bandwidth and shape factor. While quite good nose selectivity is achieved with simple ceramic filters, multiple resonators are required for good shape factors (see Fig 6.21). Some filters are of 'hybrid' form using combinations of inductors and ceramic resonators.

Examples of ceramic filters include the Philips LP1175 in which a hybrid unit provides the degree of selectivity associated with much larger conventional IF transformers; a somewhat similar arrangement is used in the smaller Toko filters such as the CFT455C which has a bandwidth (to -6dB) of 6kHz. A more complex 15-element filter is the Murata CFS-455A with a bandwidth of 3kHz at -6dB, 7.5kHz at -70dB and insertion loss 9dB, with input and output impedances of 2kΩ and centre frequency of 455kHz. In general ceramic filters are available from 50kHz to about 10.7MHz centre frequencies. Fig 6.42 provides more details.

Ceramic filters tend to be more economical than crystal or mechanical filters but have lower temperature stability and may have greater pass-band attenuation.

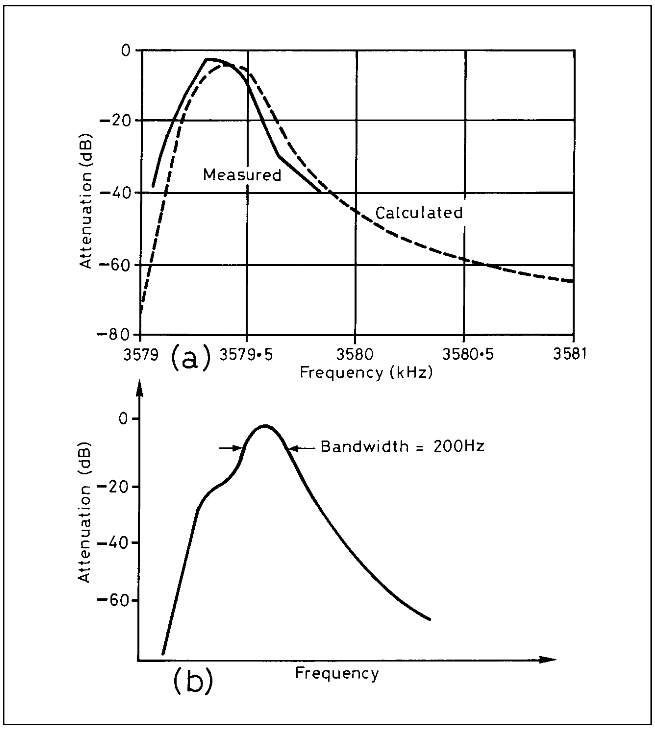


Fig 6.41: (a) Calculated and measured response curves for the CW filter. (b) Filter response resulting from use of poorly matched crystals

Mechanical and Miscellaneous Filters

Very effective SSB and CW filters at intermediate frequencies from about 60 to 600kHz depended on the mechanical resonances of a series of small elements usually in the form of discs.

The mechanical filter consisted of three basic elements: two magneto-striction transducers which convert the IF signals into mechanical vibrations and vice versa; a series of metal discs mechanically resonated to the required frequency; and disc coupling rods. Each disc represents a high-Q series resonant circuit and the bandwidth of the filter is determined by the coupling

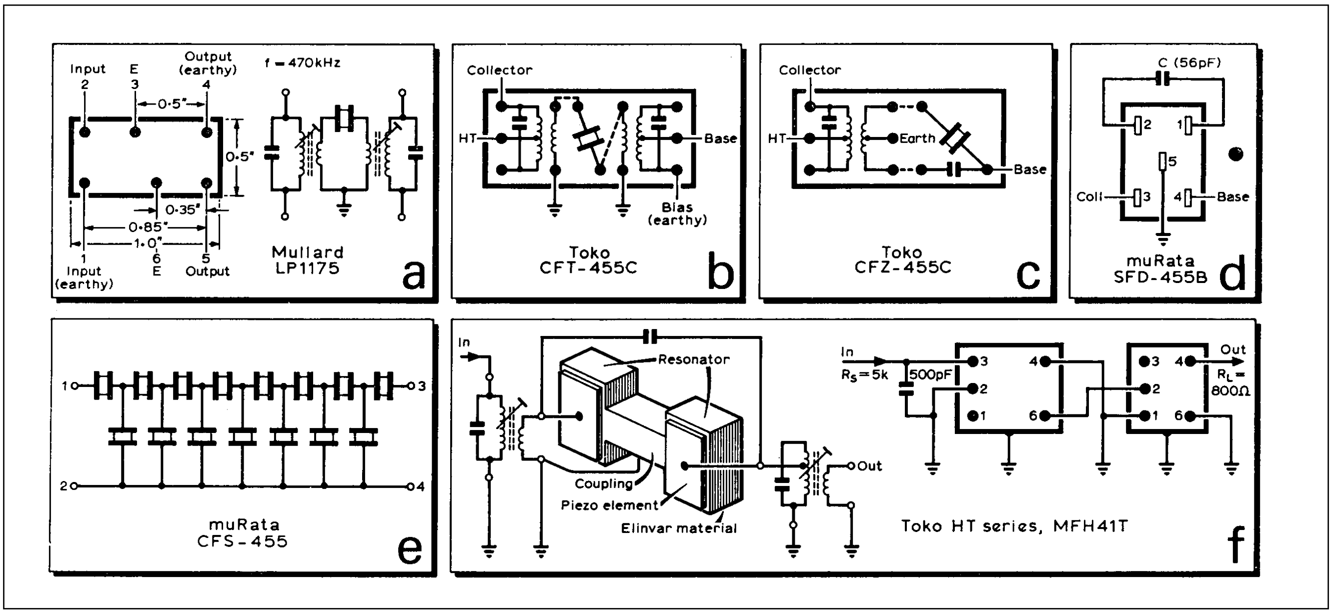


Fig 6.42: Representative types of ceramic filters

rods. 6-60dB shape factors can be as low as about 1.2, with low pass-band attenuation. The limitation of mechanical filters to frequencies of about 500kHz or below has led to their virtual disappearance from the amateur markets and they are now found only in older models despite their excellent performance below 500kHz.

Other forms of mechanical filters have been developed which include ceramic piezoelectric transducers with mechanical coupling: they thus represent a combination of ceramic and mechanical techniques. These filters may consist of an H-shaped form of construction; such filters include a range manufactured by the Toko company of Japan. Generally the performance of such filters is below that of the disc resonator type, but can still be useful.

Surface acoustic wave (SAW) filters are available for possible band-pass filter applications where discrete-element filters have previously been used, including IF filters. Filters in the 80 to 150MHz region are manufactured for cellular telephone applications, and those designed for the American AMPS and IS136 standards may be narrow enough (<30kHz) to be useful as roofing filters. It should be noted that the mechanism of signal propagation in the SAW filter is such that they are extremely resistant to intermodulation difficulties.

CIRCUITRY

Receiver Protection

Receivers, particularly where they are to be used alongside a medium- or high-powered transmitter, need to be protected from high transient or other voltages induced by the local transmitter or by build-up of static voltages on the antenna. Semiconductors used in the first stage of a receiver are particularly vulnerable and invariably require protection.

The simplest form of protection is the use of two diodes in back-to-back configuration. Such a combination passes signals less than the potential hill of the diodes (about 0.3V for germanium diodes, about 0.6V for silicon diodes) but provides virtually a short-circuit for higher-voltage signals. This system is usually effective but has the disadvantage that it introduces non-linear devices into the signal path and may occasionally be the cause of cross- and inter-modulation.

The MOSFET devices are particularly vulnerable to static puncture and some types include built-in zener diodes to protect the 'gates' of the main structure. Since these have limited rating it may still be advisable to support them with external diodes or small gas-filled transient suppressors.

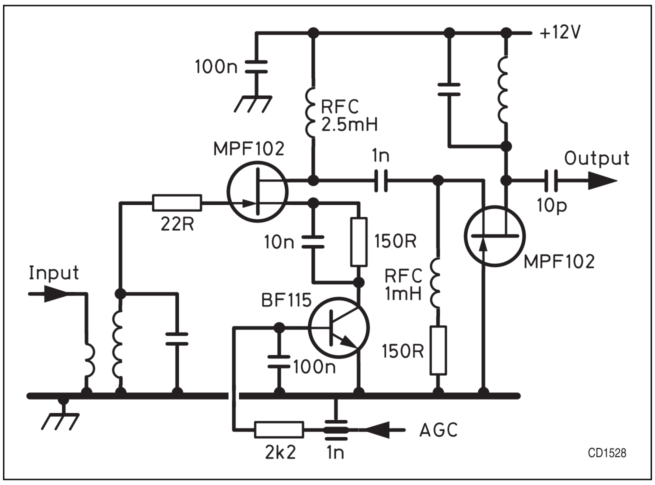


Fig 6.43: Cascode RF amp using two JFETs with transistor as AGC

Input Circuits and RF Amplifiers

It has already been noted that with low-noise mixers it is now possible to dispense with high-gain RF amplification. Amplifiers at the signal frequency may, however, still be advisable to provide: pre-mixer selectivity; an AGC-controlled stage which is in effect a controlled attenuator on strong signals; to counter the effects of conversion loss in diode and FET-array mixer stages.

In practice semiconductor RF stages are often based on junction FETs as shown in **Fig 6.43** or dual-gate MOSFETs, or alternatively integrated circuits in which large numbers of bipolar transistors are used in configurations designed to increase their signal-handling capabilities.

Tuned circuits between the antenna and the first stage (mixer or RF amplifier) have two main functions: to provide high attenuation at the image frequency; to reduce as far as possible the amplitude of all signals outside the IF pass-band.

Coupled tuned circuits

Most amateur receivers still require good pre-mixer selectivity; this can be achieved by using a number of tuned circuits coupled through low-gain amplifiers, or alternatively by tuneable or fixed band-pass filters that attenuate all signals outside the amateur bands (**Fig 6.44**). The most commonly used input arrangement consists of two tuned circuits with screening between them and either top-coupled through a small-value fixed capacitor, or bottom-coupled through a small common inductance.

In order that the coupling is maintained constant over the tuning range, the coupling element should be the same sort as the fixed element, as in **Fig 6.44(b)**. Critical coupling is achieved when the coupling coefficient, k, multiplied by the working Q is equal to 1. For **Fig 6.44(a)**, k is given by:

$$k = Cc/\sqrt{C_1C_2}$$

where C_1, C_2 are the tuning capacitors, and Cc is the top coupling capacitor. In Fig 6.44(b):

$$k = Lc/\sqrt{L_1L_2}$$

where L_1, L_2 are the tuning inductors, and Lc is the coupling inductor.

Before integrated circuits, receiver IF amplifiers were built up with a chain of valves or transistors coupled together with transformers. Each transformer was a usually a double tuned circuit

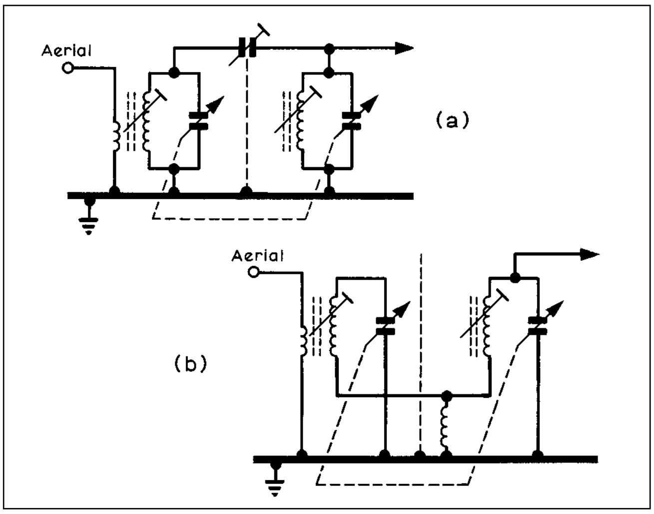


Fig 6.44: Typical RF input circuits used to enhance RF selectivity and capable of providing more than 40dB attenuation of unwanted signals 10% off tune

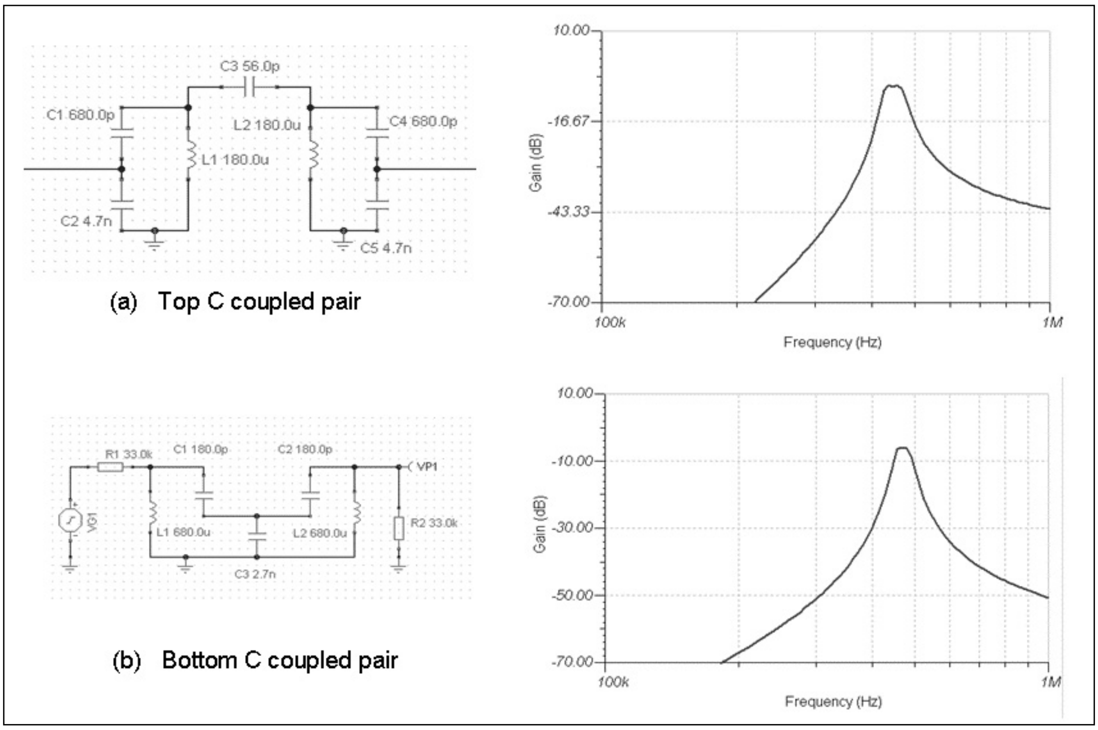


Fig 6.45: Two methods of connecting tuned circuits

providing a proportion of the overall selectivity.

With the advent of integrated circuits, all of the IF gain became a single block with the selectivity now in front between the mixer and the input to the IF amplifier. This was good because of the protection given to the IF amplifier against strong off channel signals. However, the detector now experienced all the wideband noise generated within the IF amplifier.

Ideally the overall IF selectivity should be provided by both input filtering where most of the selectivity takes place, plus output filtering to eliminate the wideband noise generated within the IF amplifier.

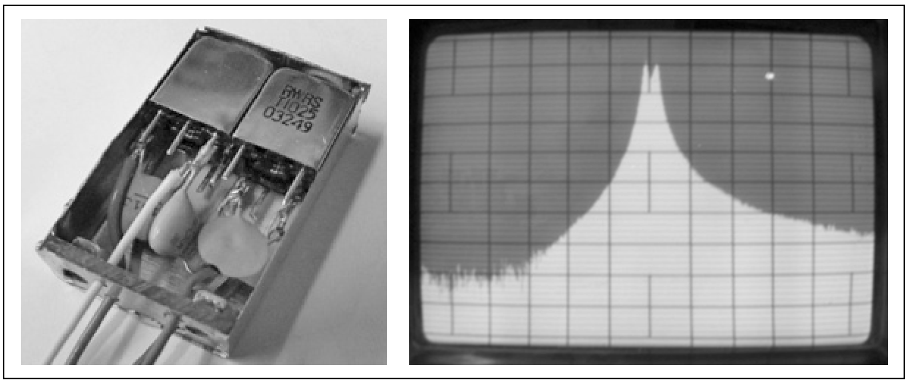


Fig 6.46: (left) Assembling two TOKO coils into a 455kHz transformer; (right) The resulting passband [50KHz per div and 10dB per div]

The Principles chapter introduces the coupled tuned circuit IF transformer, and the chapter on Passive Components illustrates its practical implementation. In essence, there are two tuned circuits whose physical spacing determines the coupling and hence the bandwidth of the transformer. There are several other ways of providing the all important coupling, two of which are illustrated in **Fig 6.45**.

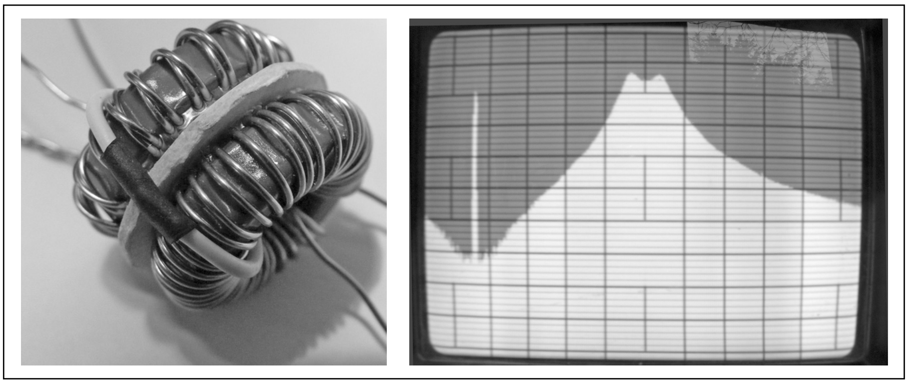


Fig 6.46: (left) Construction of coupled pair using iron dust toroids. Note the single turn inductive coupling; (right) Passband centred on 5.5MHz [1MHz per div and 10dB per div]

Top C coupling is where a small value capacitor connects the two hot ends of the identical tuned circuits, as illustrated in Fig 6.45(a). A feature of top C coupling is that the attenuation falls away more rapidly on the low frequency side of resonance. Bottom C coupling is achieved by connecting the cold ends of the two coils together and then shunting away most of the energy by use of a large value capacitor as shown in Fig 6.45(b). Bottom C coupling is characterised by the attenuation falling off more rapidly on the high side of resonance. One advantage of bottom C coupling is if the two halves of the pair are physically separated, they can be joined via a piece of coax that forms part of the bottom C.

Whilst the traditional type of IF transformer can still be found at rallies, they are too big for modern construction practice. Most home built receivers will use the Toko 10mm 10K and 10EZ range of coils that are out of production but still available from a few stockists such as [10]. However this range does not include double tuned transformers and so some way of fabricating the double tuned device is required. **Fig 6.46** illustrates a way of assembling an IF transformer centred on 455KHz from two Toko coils using top C coupling.

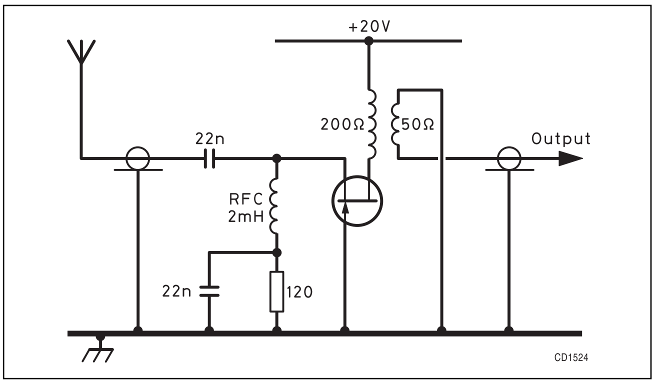


Fig 6.48: Broad-band RF amplifier using power FET and capable of handling signals to almost 3V p-p, 0.5 to 40MHz with 2.5dB noise figure and 140dB dynamic range. Drain current 40mA. Voltage gain 10dB. A suitable device would be a 2N5435 FET

Also shown is the measured frequency response illustrating the characteristic dip in the middle of the passband due to slight overcoupling and the reduced attenuation on the high side of resonance. The particular TOKO coil inductance was 680µH tuned with an external 180pF capacitor and a coupling capacitor of 5.6pF.

Another form of IF transformer is shown in **Fig 6.47.** This example was designed to give a passband of 5.0 to 5.5MHz, and was assembled using two iron dust toroids with inductive coupling between them.

RF amplifiers and attenuators

Broad-band and untuned RF stages are convenient in construction but can be recommended only when the devices used in the front-end of the receiver have wide dynamic range. An example shown in **Fig 6.48** is a power FET designed specifically for this application and operated in the earthed-gate mode suitable for use on incoming low-impedance coaxial feeders.

Unless the front-end of the receiver is capable of coping with the full range of signals likely to be received, it may be useful to fit an attenuator working directly on the input signal. **Fig 6.49** shows simple techniques for providing manual attenuation control; **Fig 6.50** is a switched attenuator providing constant impedance characteristics.

A wide-band amplifier placed in front of a mixer of wide dynamic range must itself have good spurious free dynamic range. Dynamic range is defined as the ratio of the minimum detectable signal (equal to the noise floor of the receiver, ie 3dB signal-plus-noise-to-noise ratio) to that signal level which leads to the intermodulation products being equal to the noise floor (ie 3dB signal-plus-noise-to-noise ratio).

Amplifiers based on power FETs can approach 140dB dynamic range when operated at low gain (about 10dB) and with about 40mA drain current. This compares with about 90 to 100dB for good valves (eg E810F), 80 to 85dB for small-signal FET devices; and 70 to 90dB for small-signal bipolar transistors. A high-dynamic-range bipolar amplifier can exhibit a third-order input intercept point of +33dBm with a noise figure of 3dB, using suitable transistors and noiseless feedback techniques, representing about 160dB of dynamic range.

Fig 6.51 shows two wide-band RF amplifiers, both capable of better than +40dBm OIP3.

The dynamic range of an amplifier can be increased by the operation of two devices in a balanced (push-pull) mode, but at a penalty of a 3dB increase in noise figure.

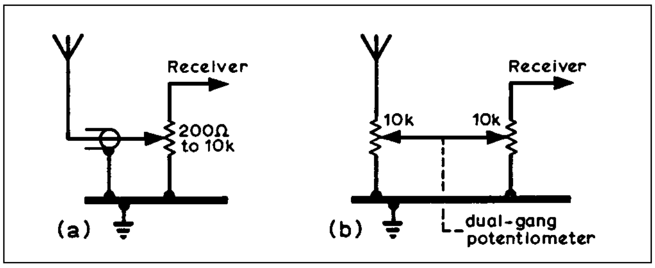


Fig 6.49: Simple attenuators for use in front of a receiver of restricted dynamic range. (a) No attempt is made to maintain constant impedance. (b) Represents less change in impedance

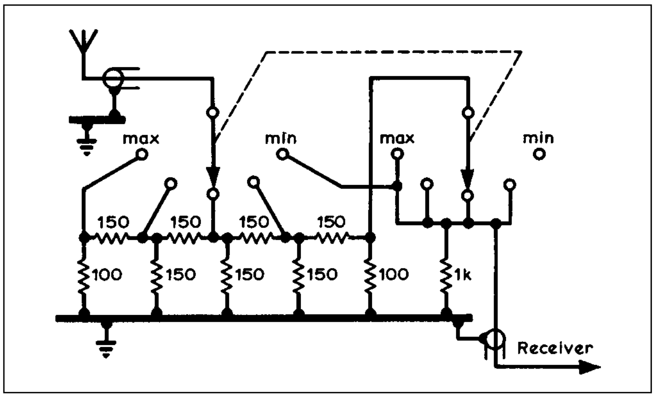


Fig 6.50: Switched antenna attenuator for incorporation in a receiver

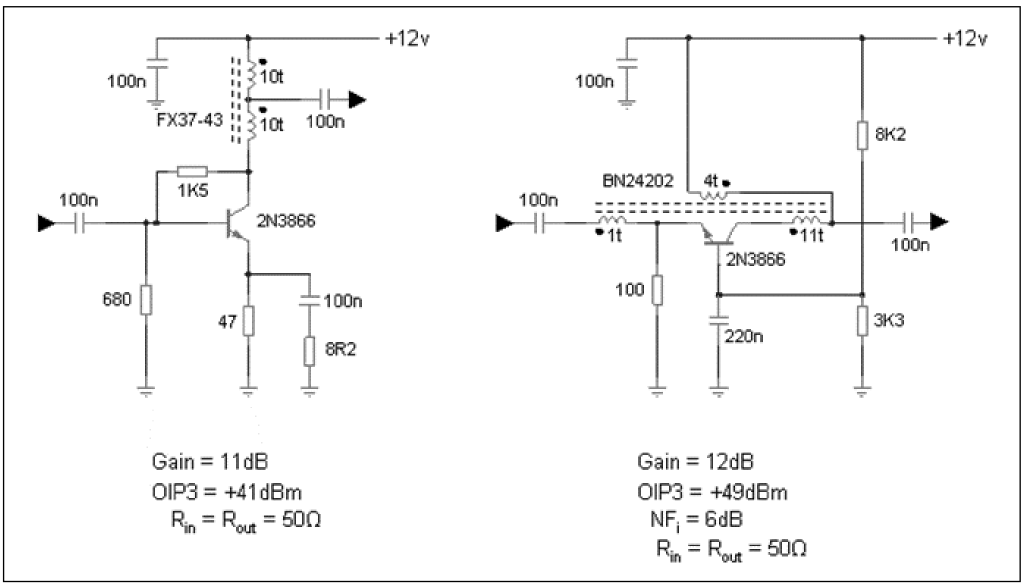


Fig 6.51: Two wideband RF amplifiers

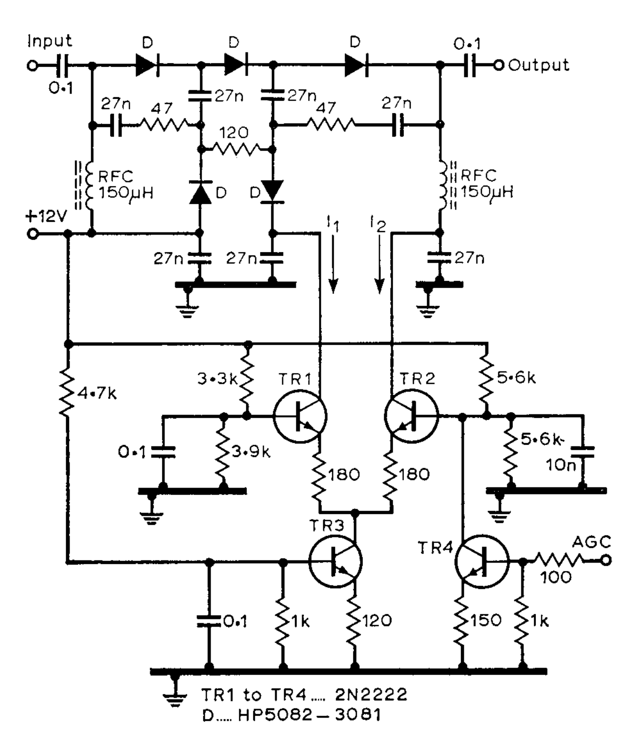


Fig 6.52: Five PIN diodes in a double-T arrangement form an AGC-controlled attenuator. The sum of the transistor collector currents is maintained constant to keep input and output impedances constant

Various forms of attenuators controlled from the AGC line are possible. Fig 6.52 shows a system based on PIN diodes; Fig 6.53 is based on toroid ferrite cores and can provide up to about 45dB attenuation when controlled by a potentiometer. Fig 6.54 shows an adaptation with MOSFET control element for use on AGC lines, although the range is limited to about 20dB.

The tuned circuits used in front-ends may be based on toroid cores since these can be used without screening with little risk of oscillation due to mutual coupling. However, the available Q cannot always compete with that of air-cored coils of thick wire, or of bunched conductor wire and pot cores at the lower frequencies.

It is important to check filters and tuned circuits for non-linearity in iron or ferrite materials. Intermodulation can be caused by the cores where the flux level rises above the point at which saturation effects begin to occur. This is not usually a problem with dust iron cores, however.

Fig 6.48 shows an amplifier with a dynamic range approaching 140dB and suitable for use in front of, or immediately behind, a double-balanced Schottky diode mixer. It should be appreciated, however, that power FETs are relatively expensive devices, although there are some lower-cost devices such as the J310.

Since the optimum dynamic range of an amplifier is usually achieved when the device is operated at a specific working point (ie bias potential) it may be an advantage to design the stage for fixed (low) gain with front-end gain controlled by means of an antenna attenuator (manual or AGC-controlled). Attenuation of signals ahead of a stage subject to cross-modulation is often beneficial since 1dB of attenuation reduces cross-modulation by 2dB.

For semiconductor stages using FETs and bipolar transistors the grounded-gate and grounded-base configuration is to be preferred. Special types of bipolar transistors (such as the

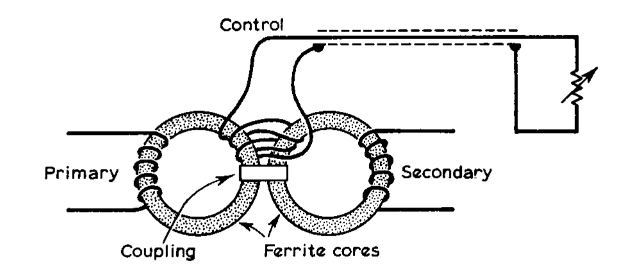


Fig 6.53: Basic form of RF level control using two toroidal ferrite cores

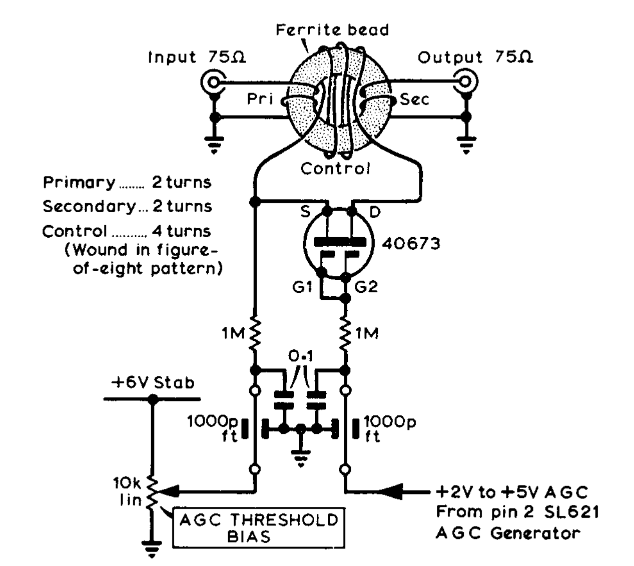


Fig 6.54: Automatic antenna attenuator based on the technique shown in Fig 6.53

BF314, BF324 developed for FM radio tuners) provide a dynamic range comparable with many junction FETs, a noise figure of about 4dB and a gain of about 15dB with a collector current of about 5mA.

Generally, the higher the input power of the device, the greater is likely to be its signal-handling capabilities: overlay and multi-emitter RF power transistors or those developed for CATV applications can have very good intermodulation characteristics.

For general-purpose, small-signal, un-neutralised amplifiers the zener-protected dual-gate MOSFET is probably the best and most versatile of the low-cost discrete semiconductor devices, with its inherent cascode configuration. If gate 2 is biased initially at about 30 to 40% of the drain voltage, gain can be reduced (manually or by AGC action) by lowering this gate 2 voltage, with the advantage that this then increases the signal-handling capacity at this stage. This type of device can be used effectively for RF, mixer, IF, product detector, AF and oscillator applications in HF receivers.

Mixers

In the overall evolution of HF receivers sensitivity, selectivity and frequency stability have been mastered. Further improvements in receiver performance are now dependent on their ability to handle large signals ie good linearity. Consequently much attention has been given in recent years to improving mixer performance in order to make superhet designs less subject to spurious responses and to improve their ability to handle weak signals in the presence of strong unwanted signals. In particular there has been increasing use of low-noise balanced and double-balanced mixers, sometimes constructed in wide-band form. Ideally mixers are simply multipliers, as are product detectors.

Mixers operate either in the form of switching mixers (the normal arrangement with diode mixers) or in what are termed continuous non-linear (CNL) modes. Generally, switching mixers can provide better performance than CNL modes but require more oscillator injection, preferably in near-square-wave form. The concept of 'linearity' in mixers may seem a contradiction in terms, since in order to introduce frequency conversion the device must behave in a highly non-linear fashion in so far as the oscillator/signal mixing process is concerned, and the term 'linearity' refers only to the signal path.

Because of their near square-law characteristics field-effect devices make successful mixers provided that care is taken on the oscillator drive level and the operating point (ie bias resistor); preferably both these should be adjusted to suit the individual device used. Even so, they are not recommended for high-performance receivers except as a balanced mixer.

Optimum performance of a mixer requires correct levels of the injected local oscillator signal and operation of the device at the correct working point. This is particularly important for FET devices. Some switching-mode mixers require appreciable oscillator power.

Junction FETs used as mixers can be operated in three ways:

(a) RF signal applied to gate, oscillator signal to source;
(b) RF signal to source, oscillator signal to gate; and
(c) RF and oscillator signals applied to gate.

Approach (a) provides high conversion gain but requires high oscillator power and may result in oscillator pulling; (b) gives good freedom from oscillator pulling and requires low oscillator power, but provides significantly lower gain; (c) gives fairly high gain with low oscillator power and may often be the optimum choice. For all FET mixers careful attention must be paid to operating point and local oscillator drive level.

For most applications the dual-gate MOSFET mixer (**Fig 6.55**) probably represents the best of the 'simple' single-device arrangements. However, care needs to be taken in that coupling of signals from gate 2 to gate 1 can lead to undesirable radiation and in some cases, difficult-to-find spurious responses.

During the 1980s and '90s, much attention was given to improving the input intercept point of receivers in order to cope better with the range of strong signals that reach low-noise broad-band mixers with a minimum of pre-mixer selectivity. A number of high-performance, double-balanced mixers were developed which were available in packaged form for home-construction.

It has been noted that a double-balanced mixer offers advantages over single-device and single-balanced mixers in reducing the number of IMPs, and also the oscillator radiation from the antenna and the oscillator noise entering the IF channel.

Double-balanced mixers using four hot-carrier diodes (eg the ones shown in **Fig 6.56 and 6.57**) when driven at a suitable level and with correct output impedance can provide third-order intercept points (IP3) of up to about +40dBm, representing a useful margin over that required for high-performance amateur operation (about +20-25dBm), a figure that is more likely to be achieved with packaged DBMs at reasonable cost. Conversion loss is likely to be at least 6dB, and optimum performance requires a high-level of oscillator injection (up to about +20-30dBm) with a near-square waveform.

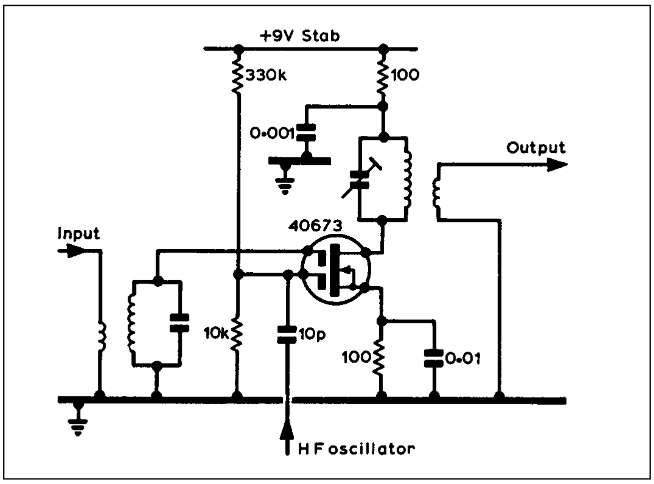


Fig 6.55: Typical dual-gate MOSFET mixer - one of the best 'simple' semiconductor mixers providing gain and requiring only low oscillator injection, but of rather limited dynamic range

An active form of double-balanced ring modulator can be based on four symmetrical medium-power FETs. Such mixers are more sensitive than diodes to changes in termination, particularly reactive components.

A less costly approach is the use of a push-pull arrangement with N-junction FETs or dual-gate MOSFETs (which provide slightly higher conversion gain) Although the gates of MOSFETs do not consume power, they do require an appreciable voltage swing in order for the mixer to operate correctly, owing to the considerable signal voltage across the FET switch and the signal current through it unless the mixer is configured so that these signal currents and voltages do not appear across the FETs.

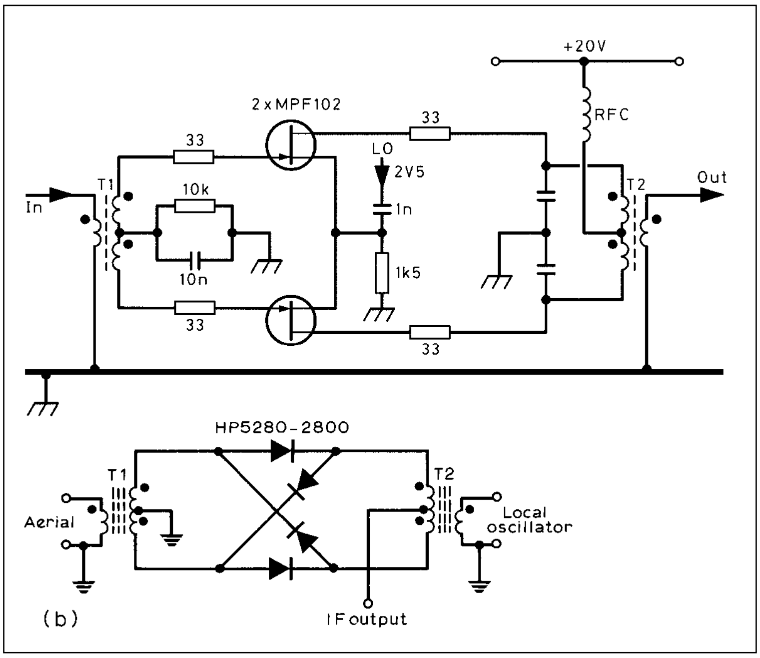


Fig 6.56: Low-noise mixers of wide dynamic range. (a) Balanced FET mixer (preferably used with devices taking fairly high current). (b) Diode ring mixer using Schottky (hot-carrier) diodes

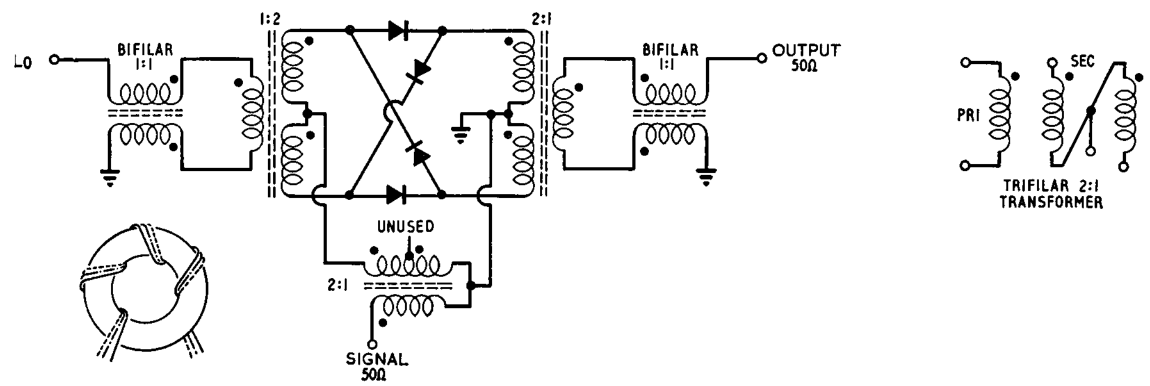


Fig 6.57: Double-balanced diode ring mixer showing how additional bifilar-wound transformers can be added to improve balance, with details of the transformers. The three strands of wire should be twisted together before winding; each winding consists of 12 to 20 turns (depending on frequency range) of No 32 enamelled wire. Injection signal should be 0.8 to 3V across 50 ohms (4-12mW). The toroid material is dependent on the frequency range

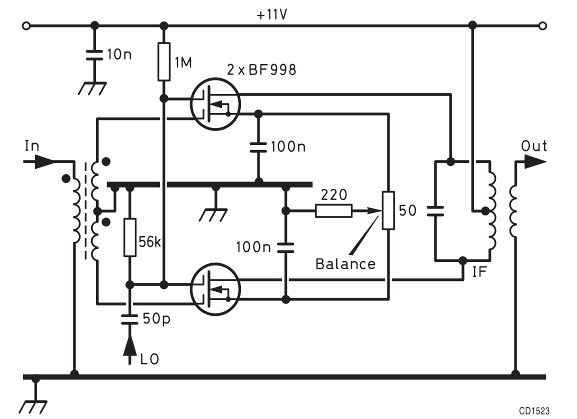


Fig 6.58: Balanced mixer using dual-gate MOSFETs

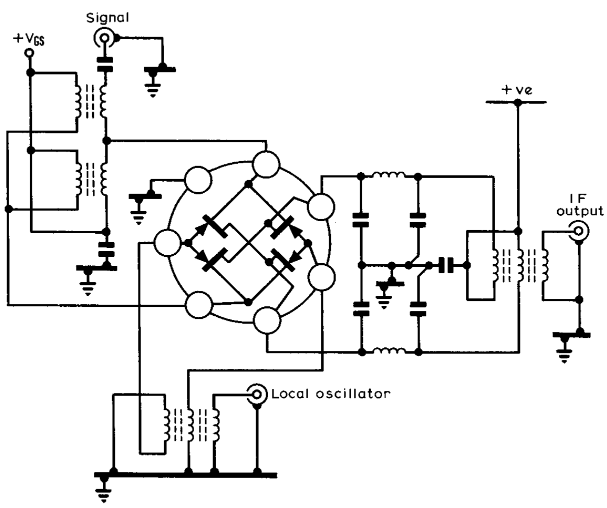


Fig 6.59: Double-balanced active FET mixer of wide dynamic range using JFET quad of power-type FETs

An ultra-low-distortion HF switched FET mixer was devised by Eric Kushnik [11] in which an IP3 of +25dBm has been measured with -3dBm local oscillator power. Some of the 74HCT series of CMOS transmission gate ICs have been successfully used in this application. FET mixers are shown in Figs 6.58 and 6.59.

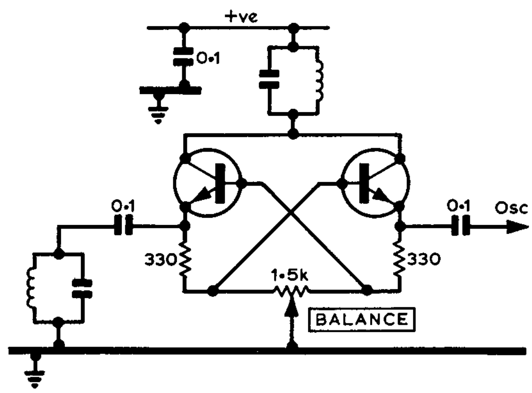


Fig 6.60: Use of cross-coupled transistors to form a double-balanced mixer without special balanced input transformers

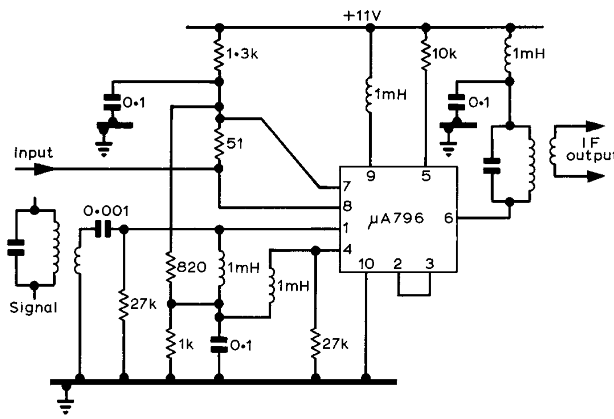
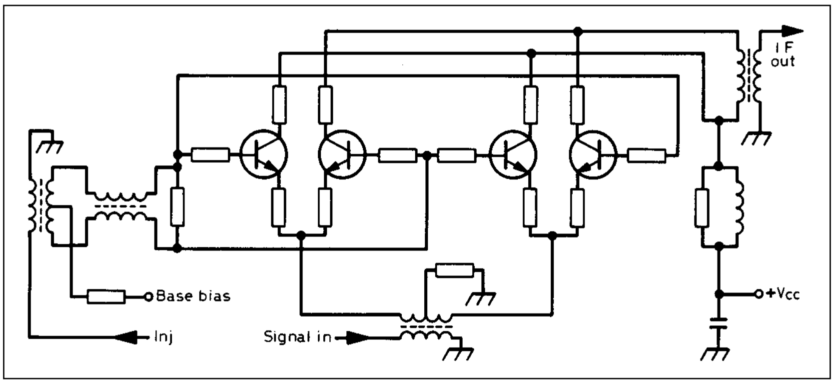


Fig 6.61: Double-balanced IC mixer circuit for use with μA796 or MC1596G etc devices

Both active and passive FET switching-mode mixers can provide a wide dynamic range; the active mixer provides some conversion gain, the passive arrangement in which the devices act basically only as switches results, as with diode mixers, in some conversion loss and must be followed by a low-noise IF amplifier, preferably with a diplexer arrangement that presents a constant impedance over a wide frequency range.

Fig 6.62: A double-balanced active mixer using bipolar transistors in a denegrated version of the balanced transconductance mixer (*Radio Receivers* - Gosling (Ed))



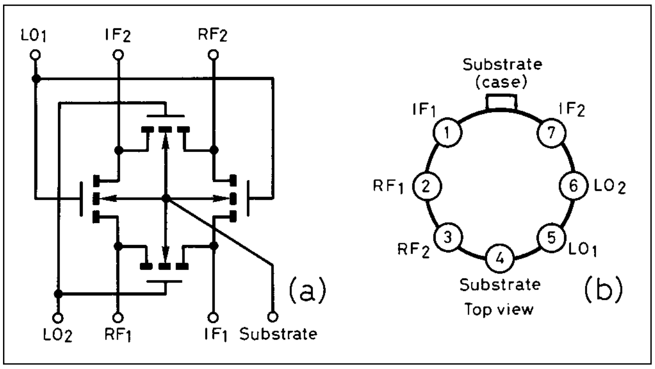


Fig 6.63: Siliconix Si8901 ring demodulator/balanced mixer. (a) Functional block diagram. (b) Pin configuration with Si8901A in TO-78 and Si8901Y as surface-mounted So14 configuration

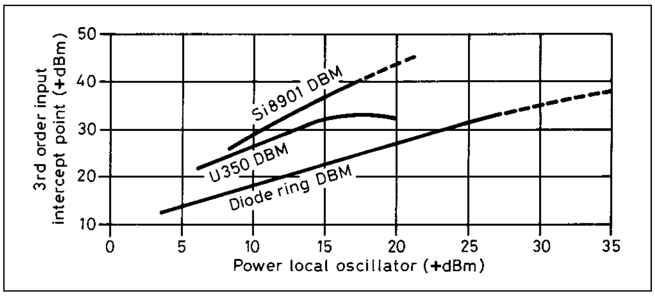


Fig 6.64: Performance comparison between Si8901 DBM, U350 active FET DBM and diode ring DBM

Characteristic	Single-ended	Single-balanced	Double-balanced
Bandwidth	Several decades	Decade	Decade
Relative intermodulation density	1	0.5	0.25
Interport isolation	Little	10-20dB	>30dB
Relative oscillator power	0dB	+3dB	+6dB

Table 6.3: Basic mixer arrangements

High-level ring-type mixers can also be based on the use of medium-power bipolar transistors in cross-coupled 'tree' arrangements (**Fig 6.60**). With such an arrangement there need be no fundamental requirement for push-pull drive or balanced input/output transformers, although their use minimises local oscillator conducted emissions. This approach is used at low-level in the Motorola MC1596 and now-obsolete Plessey SL640 IC packages, and at high-level in the obsolete Plessey SL6440. The SL6440 was specifically developed as a high-level, low-noise mixer and is capable of +30dBm intercept point, +15dBm 1dB compression point with a conversion 'gain' of -1dB. A circuit for the µA796 and similar ICs is shown in **Fig 6.61**.

Fig 6.62 shows the basic arrangement of a similar class of mixer, comprising a pair of transconductance mixers with emitter resistors added for IM improvement. Resistors in the base and collector leads add loss of ultra-high frequencies to suppress parasitic oscillations caused by resonances formed by circuit and transistor capacitances together with the leakage reactance of the associated transformers. Injection is via a balancing transformer to the bases of the bipolar transistors which are overdriven, resulting in signal switching action. Collector supply voltage is applied to the output transformer centre-tap through a parallel resistor-inductor to further suppress oscillation. With medium-power transistors, this active mixer can give a 3dB gain, 9dB noise figure and +25dBm input intercept over the 2 to 30MHz band as an up-converter to the 100MHz range. This device can be used with or without preamplification and with little pre-mixer selectivity in high-performance home-built receivers.

Common to both diode and switching-mode FET mixers is their square-law characteristics, an important factor in maintaining low distortion during mixing. Equally important for high dynamic range is the ability to withstand overload that can be a major cause of distortion. Some designs with passive ring mixers use paralleled diodes to provide greater current handling; the penalty attached to this approach is the need for a very large increase in local-oscillator power.

A form of high-performance mixer using monolithic quad-ring double-diffused MOSFETs developed by Ed Oxner, KB6QJ [12] used a resonant-gate drive transformer to provide sufficient switching voltage without a corresponding increase in switching power, an effective technique but not easily implemented with broad-band mixers.

Switching mixers, if they are to achieve a high third-order intercept point, require a drive that must:

- approach ideal square wave;
- ensure a 50% duty cycle; and
- have sufficient amplitude to switch the devices fully 'on' and 'off', and, in the case of FET devices, offer minimum resistance when 'on'.

Further, to maintain good overall performance in terms of minimum conversion loss, maximum dynamic range (ie taking into account the noise figure) and maximum strong signal performance, it is desirable to incorporate image-frequency termination. With any switching mixer operating directly on the incoming RF signals without pre-mixer amplification, the IF amplifier that follows the mixer, either directly or after a roofing filter, must have a low noise figure and a high intercept figure, preferably with a

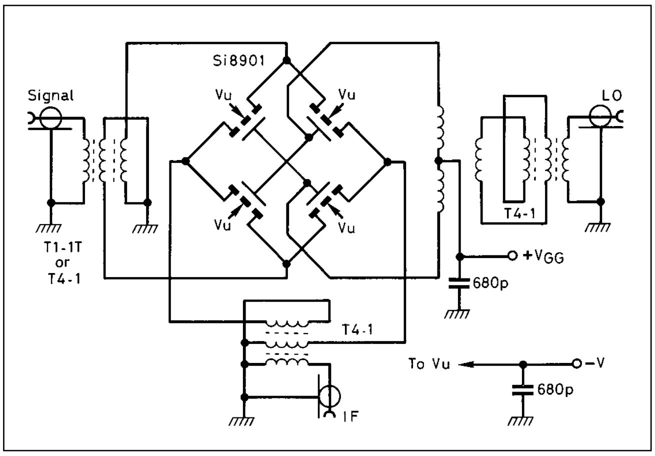


Fig 6.65: Prototype commutation double-balanced mixer as described in the Siliconix applications notes

diplexer arrangement to achieve constant input impedance over a broad band of frequencies.

Details of Siliconix mixers and their performance are shown in **Figs 6.63 to 6.65**.

Image-rejection mixer

In recent years, there has been a marked trend towards multi-conversion general-coverage receivers with the first IF in the VHF range since this facilitates the use of frequency synthesisers covering a single span of frequencies with a broad-band up-conversion mixer. It has long been considered that subsequent down-conversion mixer stages should not change the frequency by a factor of more than about 10:1 in order to minimise the 'image' response. Thus a receiver with a first IF of the order of 70MHz or 45MHz and a final IF of, say, 455kHz or 50kHz (to take advantage of digital signal processing) normally requires an intermediate IF of, say, 9 or 10.7MHz and possibly a further IF of about 1MHz. With triple or even quadruple conversions, it becomes increasingly difficult to achieve a design free of spurious responses and of wide dynamic range.

It is possible to eliminate mid-IF stages by the use of an image-rejecting two-phase mixer akin to the form of demodulator used in two-phase direct conversion receivers; this is best implemented using two double-balanced diode-ring mixers or the equivalent.

Fig 6.66 outlines the basic arrangement of an image-rejection mixer. Image rejection mixers offer slightly better intermodulation performance, all other considerations being equal, because the input signal level is reduced by 3dB. This reduces the absolute power of the 3rd order intermodulation products by 9dB, and summing the products increases this level by a maximum of 3dB. However, there is a 3dB greater noise figure, and so the absolute improvement in dynamic range disappears.

Device	Advantages	Disadvantages
Bipolar transistor	Low noise figure High gain Low DC power	High intermodulation Easily overloaded Subject to burnout
Diode	Low noise figure High power handling High burn-out level	High LO drive Interface to IF Conversion loss
JFET	Low noise figure Conversion gain Excellent in performance Square-law characteristic Excellent overload High burn-out level	Optimum conversion gain not possible at optimum square law response High LO power
Dual-gate MOSFET	Low IM distortion AGC Square-law characteristic	High noise figure Poor burn-out level Unstable

Table 6.4: Device comparisons

Image rejection mixers are generally limited to about 30dB rejection: with careful trimming, possibly 40dB may be attained over time and temperature variations. Such a mixer can convert a VHF signal directly to, say, 50kHz while maintaining image response at a low level.

More about mixers can be found in the one of the Building Blocks chapters earlier in this book.

The HF Oscillator(s)

The frequency to which a superhet or direct-conversion receiver responds is governed not by the input signal frequency circuits but by the output of the local oscillator. Any frequency variations or drift of the oscillator are reflected in variation of the received signal; for SSB reception variations of more than about 50Hz will render the signal unintelligible unless the set is retuned.

Much more on this topic can be found in the chapter Building Blocks: Oscillators.

Mentioned earlier in this chapter was the unwanted effect known as reciprocal mixing which is directly related to oscillator noise. This aspect represents the next evolutionary step in HF receiver design.

IF Amplifiers

The IF remains the heart of a superhet receiver, for it is in this section that virtually all of the voltage gain of the signal and the selectivity response are achieved. Whereas with older superhets (which had significant front-end gain) the IF gain was of the order of 70-80dB, today it is often over 100dB.

Where the output from the mixer is low (possibly less than 1μV) it is essential that the first stage of the IF section should have low-noise characteristics and yet not be easily overloaded. Although it is desirable that the crystal filter (or roofing filter) should be placed immediately after the mixer, the very low output of diode and passive FET mixers may require that a stage of amplification takes place before the signal suffers the insertion loss of the filter.

Similarly it is important that where the signal passes through the sideband filter at very low levels the subsequent IF amplifier must have good noise characteristics. Further, for optimum CW reception, it will often be necessary to ensure that the noise bandwidth of the IF amplifier after the filter is kept narrow. The noise band-

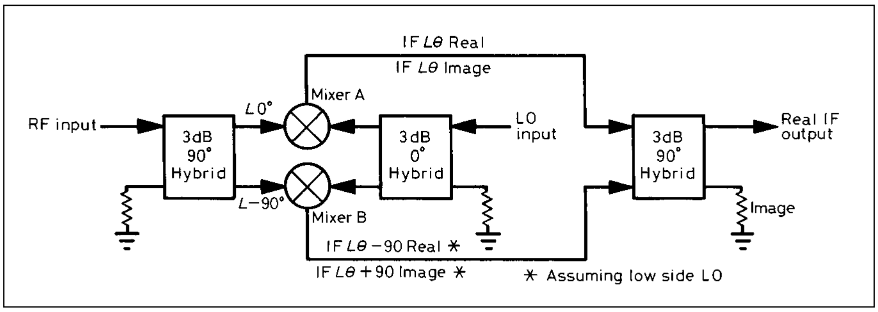


Fig 6.66: Image rejecting mixer from *Microwave Solid-state Circuit Design* by Bahl and Bhartia (1988)

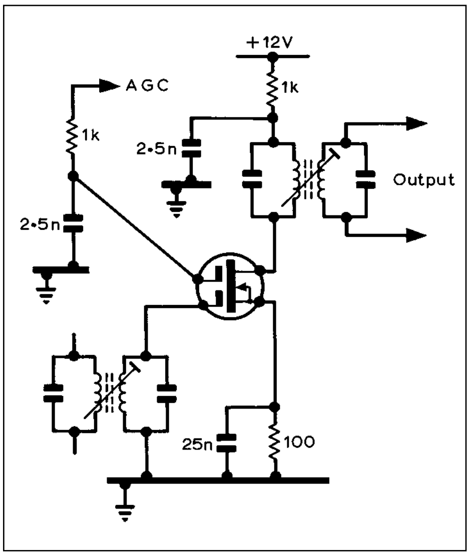


Fig 6.67: Typical automatic gain-controlled IF amplifier a dual-gate MOSFET

width of the entire amplifier should be little more than that of the filter. This can be achieved by including a further narrow-band filter (for example a single-crystal filter with phasing control) later in the receiver, or alternatively by further frequency conversion to a low IF.

To achieve a flat AGC characteristic it may be desirable for all IF amplifiers to be controlled by the AGC loop, and it is important that amplifier distortion should be low throughout the dynamic range of the control loop.

The dual-gate MOSFET (**Fig 6.67**) with reverse AGC on gate 1 and partial forward AGC on gate 2 has excellent cross-modulation properties but the control range is limited to about 35dB per stage. Integrated circuits with high-performance gain-controlled stages are available.

Where a high-grade SSB or CW filter is incorporated it is vital to ensure that signals cannot 'leak' around the filter due to stray coupling; good screening and careful layout are needed.

In multiple-conversion receivers, it is possible to provide continuously variable selectivity by arranging to vary slightly the frequency of a later conversion oscillator so that the band-pass of the two IF channels overlap to differing degrees. For optimum results this requires that the shape factor of both sections of the IF channel should be good, so that the edges are sharply defined.

With double-tuned IF transformers, gain will be maximum when the product kQ is unity (where k is the coupling between the windings). IF transformers designed for this condition are said to be critically coupled; when the coupling is increased beyond this point (over-coupled) maximum gain occurs at two points equally spaced about the resonant frequency with a slight reduction of gain at exact resonance: this condition may be used in broadcast receivers to increase bandwidth for good-quality reception.

If the coupling factor is lowered (under-coupled) the stage gain falls but the response curve is sharpened, and this may be useful in communications receivers.

FETs as Small-signal Amplifiers

Field effect transistors make good small-signal RF or IF amplifiers and offer advantages when used properly. They can provide low-noise, rugged amplifiers once wired in circuits having an easy, low-resistance path between their gate(s) and earth, though they are vulnerable to electrostatic discharge when out of circuit, unless protected by an internal diode(s). However, despite being, like thermionic valves, voltage-controlled

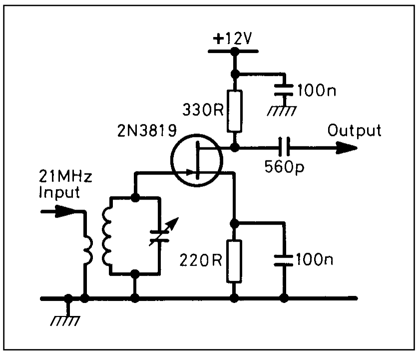


Fig 6.68: Stable FET preamp with low-impedance output load and with the main part of the gain coming from the step-up input transformer

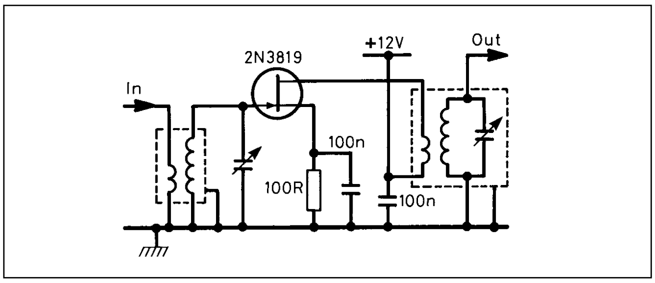


Fig 6.69: Stable FET IF amplifier using two bipolar transistor-type IF transformers in reverse configuration

devices, they should not be used as replacements for valves in similar circuits.

FETs have very high slope of up to 30mA/V (much higher than most valves) but also much greater drain-to-gate capacitance than the inter-electrode capacitances of triode valves. This is a sure prescription for self-oscillation if connected in a typical pentode-type amplifier circuit.

However, the FET can be used with a very low output load impedance and thus can be used as a stable low-gain device, yet providing excellent stage gain because of the voltage step-up that can be readily achieved with a resonant input transformer.

For example, the 21MHz preamplifier shown in **Fig 6.68** using the 2N3819 FET (10mA/V) with a 330Ω resistor as load has a device gain of only three, but the input tuned circuit can provide a voltage gain of about seven, resulting in a stable voltage gain of about 21, with the FET's very high input impedance presenting only light loading of the input transformer. This provides an unconditionally stable stage gain of over 20dB. **Figs 6.69, 6.70 and 6.71** show typical FET and dual-gate FET small-signal amplifiers.

The most critical amplifier in a modern, high-performance receiver is usually the post-mixer amplifier, ie the IF amplifier that follows the mixer, either directly or after a crystal filter. It needs to be of low noise in order to cope with the conversion loss of a ring mixer, and if preceding the crystal filter will have to

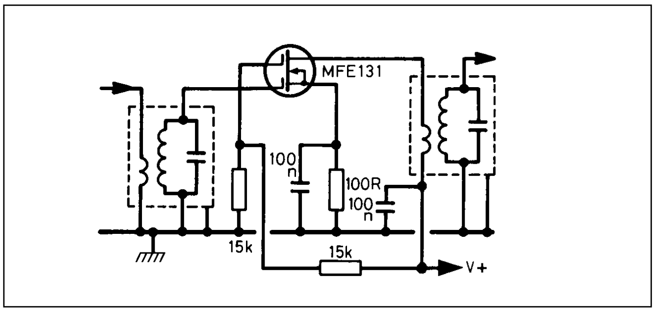


Fig 6.70: Dual-gate FET IF amplifier

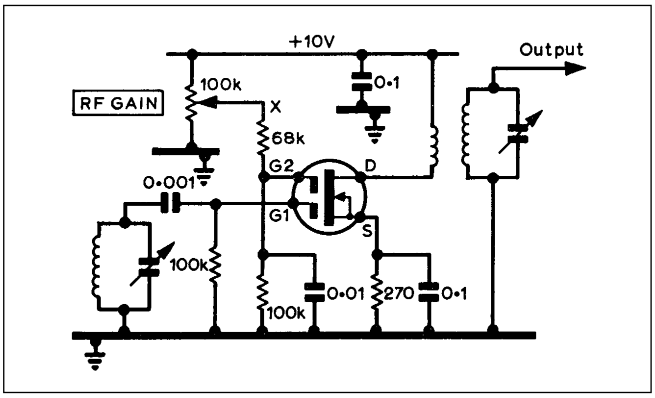


Fig 6.71: Typical dual-gate MOSFET RF or IF amplifier. G2 is normally biased to about one-third of the positive voltage of drain. In place of manual gain, control point X can be connected to a positive AGC line

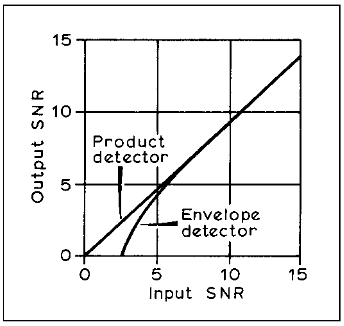


Fig 6.72: Synchronous (product) detection maintaining the SNR of signals down to the lowest levels whereas the efficiency of envelope detection falls off rapidly at low SNR, although as efficient on strong signals

cope with large off-frequency signals, requiring a high intercept characteristic. For the highest-performance receivers, push-pull bipolar transistors with a noise figure of about 2dB and a third-order intercept point equal or better than that of the mixer are required.

Possibly the simplest post-mixer amplifier for high-performance receivers is a power-FET common-gate stage, such as that shown earlier. With a 2N5435 FET this can provide a 2dB noise figure with a 50-ohm system gain of 9dB and an output third-order intercept point of +30dBm when biased at V_{DD} of 12-15V at 50mA.

Amplifier Stability Criteria

For any amplifying device, a consideration of its stability may be achieved by the application of a stability criterion. The Linville stability criterion C provides an equation for C such that when C < 1, the circuit is unconditionally stable. C is given by:

$$C = (y_{rs} \cdot y_{fs})/(2g_{is}\, g_{os} - R(y_{rs}y_{fs})$$

where y_{rs} is the magnitude of reverse transmission; y_{fs} is the magnitude of forward transmission; g_{is} is the input conductance; and g_{os} is the output conductance. $R(y_{rs}y_{fs})$ means the real part of the product of $y_{rs}y_{fs}$.

Where the characteristics of a device are given in terms of 's' or other parameters (eg 'h' parameters), a conversion can be made. The Motorola RF transistor data handbooks are particularly recommended to provide the conversion formulas between the various sets of parameters: there are also a number of standard student textbooks on the subject. The home-constructor, faced with an apparently intractable problem of stability in an RF amplifier, may well find that a few minutes spent with the device data sheet and a calculator can provide not only the explanation of the instability but also a cure. Because of this, and being based on engineering, rather than 'green fingered' principles, it is far more likely to be more repeatable and stable in the long term than a guesswork approach.

Demodulation

For many years, the standard form of demodulation for communications receivers, as for broadcast receivers, was the envelope detector using diodes. Envelope detection is a non-linear process (part mixing, part rectification) and is inefficient at very low signal levels. On weak signals this form of detector distorts or may even lose the intelligence signals altogether. On the other hand, synchronous or product detection preserves the signal-to-noise ratio, enabling post-detector signal processing and audio filters to be used effectively (**Fig 6.72**).

Synchronous detection is essentially a frequency conversion process and the circuits used are similar to those used in mixer

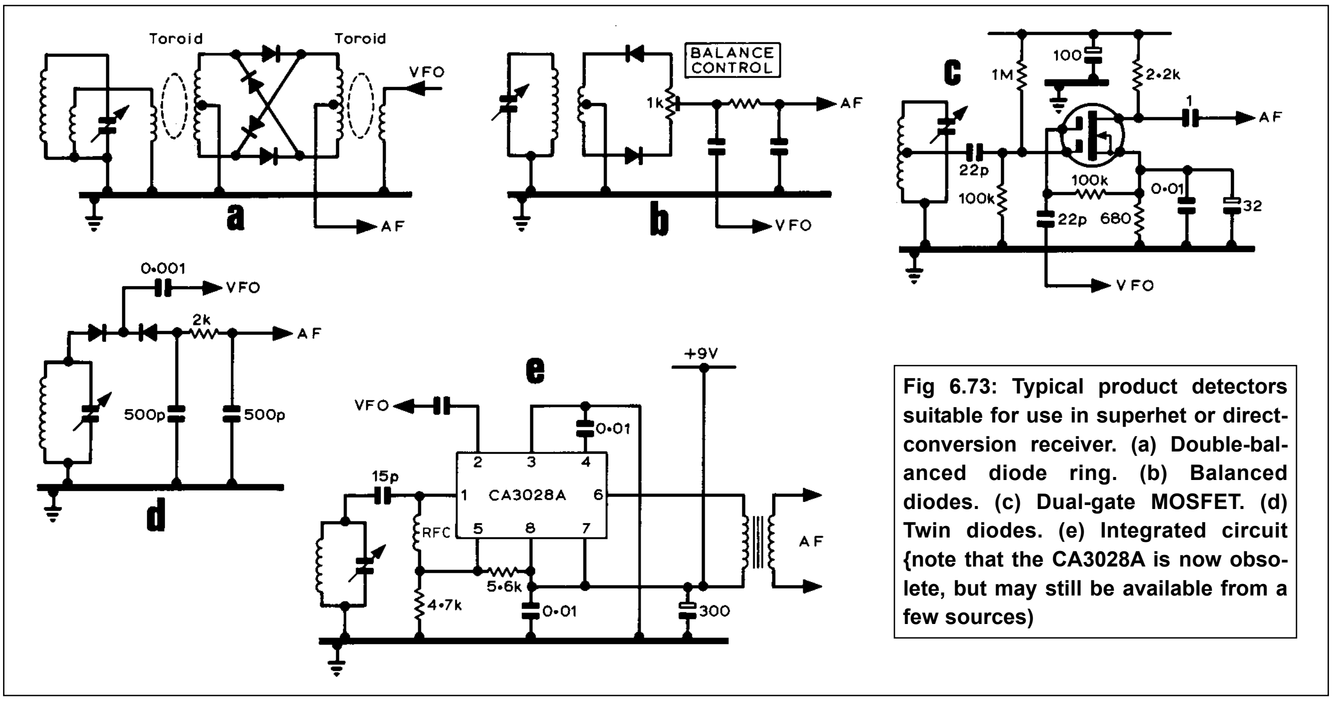


Fig 6.73: Typical product detectors suitable for use in superhet or direct-conversion receiver. (a) Double-balanced diode ring. (b) Balanced diodes. (c) Dual-gate MOSFET. (d) Twin diodes. (e) Integrated circuit {note that the CA3028A is now obsolete, but may still be available from a few sources)

stages. The IF or RF signal is heterodyned by a carrier at the same frequency as the original carrier frequency and so reverts back to the original audio modulation frequencies (or is shifted from these frequencies by any difference between the inserted carrier and the original carrier as in CW where such a shift is used to provide an audio output between about 500 and 1000Hz). Typical product detector circuits are shown in Fig 6.73.

It should be noted that a carrier is needed for both envelope and product detection: the carrier may be radiated along with the sidebands, as in AM, or locally generated and inserted in the receiver (either at RF or IF - usually at IF in superhets, at RF in direct-conversion receivers).

Synchronous or product detection has been widely adopted for SSB and CW reception in amateur receivers; the injected carrier frequency is derived from the beat frequency oscillator, which is either LC or crystal controlled. By using two crystals it is possible to provide selectable upper or lower sideband reception.

The use of synchronous detection can be extended further to cover AM, DSBSC, NBFM and RTTY but for these modes the injected carrier really needs to be identical to the original carrier, not only in frequency but also in phase: that is to say the local oscillator needs to be in phase coherence with the original carrier (an alternative technique is to provide a strong local carrier that virtually eliminates the original AM carrier - this is termed exalted carrier detection).

Phase coherence cannot be achieved between two oscillators unless some effective form of synchronisation is used. The simplest form of synchronisation is to feed a little of the original carrier into a local oscillator, so forcing a phase lock on a free-running oscillator; such a technique was used in the synchrodyne receiver. The more usual technique is to have a phase-lock loop. At one time such a system involved a large number of components and would have been regarded as too complex for most purposes; today, however, complete phase-lock loop detectors are available in the form of a single integrated circuit, both for AM and NBFM applications.

Apart from the phase-lock loop approach a number of alternative forms of synchronous multi-mode detectors have been developed. One interesting technique which synthesises a local phase coherent carrier from the incoming signal is the reciprocating detector.

Noise Limiters, Null-steerers and Blankers

The HF spectrum, particularly above 15MHz or so, is susceptible to man-made electrical impulse interference stemming from electric motors and appliances, car ignition systems, thyristor light controls, high-voltage power lines and many other causes. Static and locally generated interference from appliances, TV receivers etc can be a serious problem below about 5MHz, and may be maximum at LF/MF.

These interference signals are usually in the form of high-amplitude, short-duration pulses covering a wide spectrum of frequencies. In many urban and residential areas this man-made interference sets a limit to the usable sensitivity of receivers and may spoil the reception of even strong amateur signals.

Because the interference pulses, though of high amplitude, are often of extremely short duration, a considerable improvement can be obtained by 'slicing' off all parts of the audio signal which are significantly greater than the desired signal. This can be done by simple AF limiters such as back-to-back diodes. For AM reception more elegant noise limiters develop fast-acting biasing pulses to reduce momentarily the receiver gain during noise peaks. The ear is much less disturbed by 'holes of silence' than by peaks of noise. Many limiters of this type have been fitted in the past to AM-type receivers.

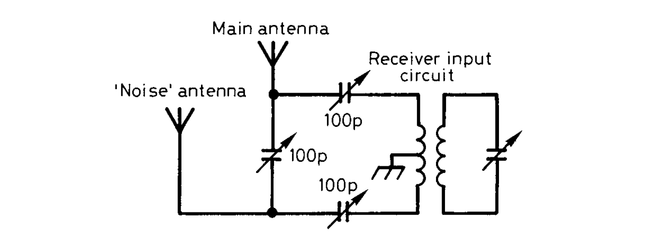


Fig 6.74: The original Jones noise-balancing arrangement as shown in early editions of *The Radio Handbook*. Local electrical interference could be phased out by means of pick-up on the auxiliary 'noise' antenna. Although it could be effective it required very careful setting up

Unfortunately, since the noise pulses contain high-frequency transients, highly selective IF filters will distort and broaden out the pulses. To overcome this problem, noise blankers have been developed which derive the blanking bias potentials from noise pulses which have not passed through the receiver's selective filters. In some cases a parallel broadly tuned receiver is used, but more often the noise signals are taken from a point early in the receiver. For example, the output from the mixer goes to two channels: the signal channel which includes a blanking control element which can rapidly reduce gain when activated; and a wide-band noise channel to detect the noise pulse and initiate the gain reduction of the signal channel. To be most effective it is necessary for the gain reduction to take place virtually at the instant that the interference pulse begins. In practice, because of the time constants involved, it is difficult to do this unless the signal channel incorporates a time delay to ensure that the gain reduction can take place simultaneously with or even just before the noise pulse. One form of time delay which has been described in the literature utilises a PAL-type glass ultrasonic television delay line to delay signals by 64 microseconds. It is, however, difficult to eliminate completely transients imposed on the incoming signal.

One possible approach, which has been investigated at the University College, Swansea, is to think in terms of receivers using synchronous demodulation at low level so that a substantial part of the selectivity, but not all of it, is obtained after demodulation. This allows noise blankers to operate at a fairly low level on AF signals. A control element which has been used successfully consists of a FET gate pulsed by signals derived from a wide-band noise amplifier. The noise gate is interposed between the mixer and the first crystal filter, with the input signal to the noise amplifier taken off directly from the mixer.

Noise limiters and noise blankers are suitable for use only on pulse-type interference in which the duty cycle of the pulse is relatively low. An alternative technique, suitable for both continuous signals and noise pulses, is to null out the unwanted signal by balancing it with anti-phase signals picked up on a short 'noise antenna'. This has led to the revival of the 1930s Jones noise-balancing technique in which local interference could be phased out by means of pick-up on an auxiliary noise antenna. This system (Fig 6.74) was capable of reducing a specific unwanted source of interference by tens of decibels while reducing the wanted signal by only about 3-6dB, but required critical adjustment of the controls.

In the early 1980s John K Webb, W1ETC, developed a more sophisticated method of phasing out interference using coiled lengths of coaxial cable as delay lines to provide the necessary phase shifts: Fig 6.75. This included a compact null-steerer located alongside the receiver/transceiver, capable of generating deep nulls against a single source of interference, resembling the nulls of an efficient MW ferrite-rod antenna. The two

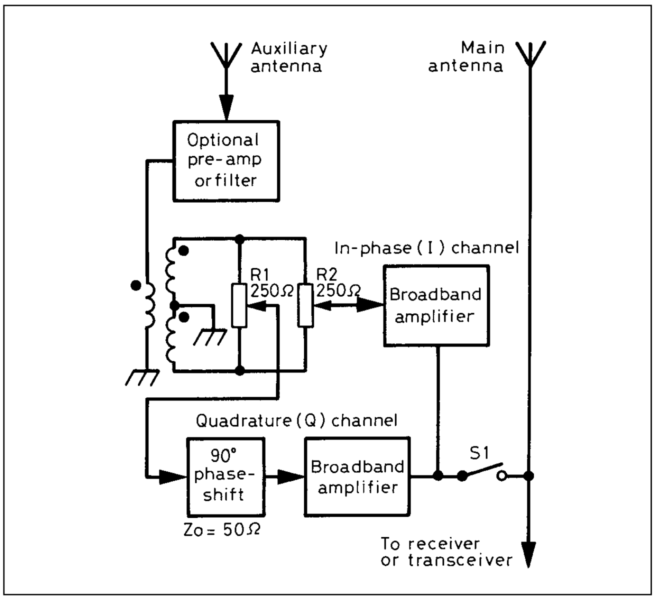


Fig 6.75: Functional diagram of the electronic null steering unit for use in conjunction with an HF receiver or transceiver (S1 is a relay contact to disconnect the system during transmission) as described by John Webb, W1ETC, of the Mitre Corporation in 1982

controls adjusted phase and amplitude of the signals from the auxiliary noise antenna. W1ETC summarised results [13] with such a unit as follows:

1. The available null depth in signals propagated over short paths of up to 20 miles is large and stable, limited only by how finely the controls are adjusted.
2. Nulls on signals arriving over short skywave paths of up to a few hundred miles are in the order of 30dB, provided there is a single mode of propagation and one direction of arrival. Such nulls are usually stable.
3. Signals propagated over paths of 10 to 100 miles may arrive as a mixture of ground-wave and skywave. A single null is thus ineffective.
4. Signals propagated by skywave over long distances frequently involve several paths, each having a different path length so that a single null has little effect on what is usually the 'wanted' signal.
5. Broad-band radiated noise can be nulled as deeply as any radio signal. This seems to be a more effective counter to noise than blanking or limiting techniques with local electrical noise deeply nulled, and with little effect on wanted long-distance signals.

To meet result (5) the interference has to be directly radiated to the receiver antennas and not enter the receiver in a less-directional manner (for example re-radiation from mains cabling etc).

Various methods of implementing null-steering have been described: **Figs 6.76-6.78** show a design by Lloyd Butler, VK5BR, utilising the phase shifts of off-tune resonant circuits.

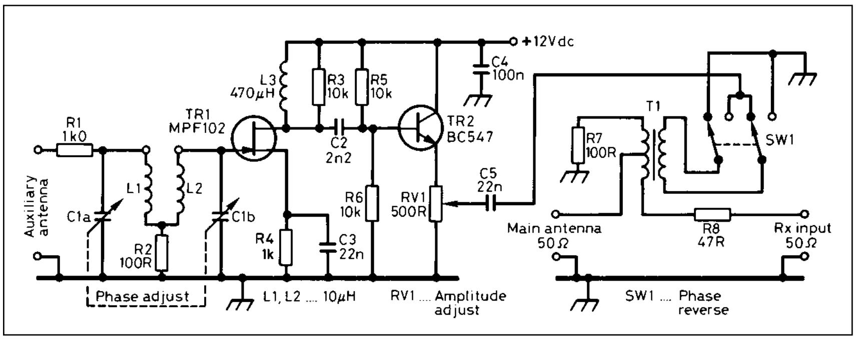


Fig 6.76: VK5BR's Mk 2 interference-cancelling circuit as described in the January 1993 issue of Amateur Radio. As shown this covers roughly 3.5 to 7MHz. C1 ganged 15-250pF variable capacitor or similar. L1, L2 miniature 10μH RF chokes. T1 11 turns quadfilar wound on Amidon FT-50-75m toroidal core

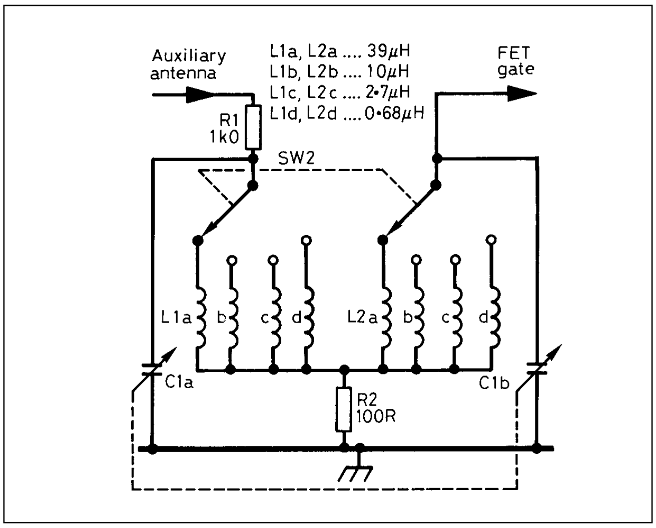


Fig 6.77: Bandswitching modification to provide 1.8 to 30MHz in four ranges

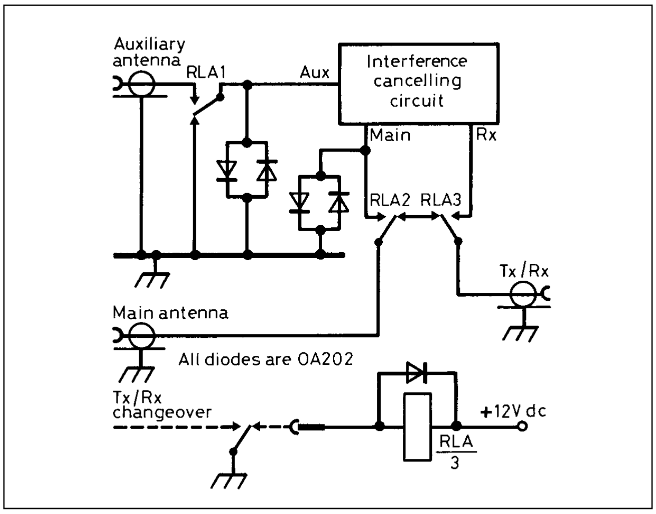


Fig 6.78: Transmit-receive switching with protection diodes for use with the VK5BR interference-cancelling circuit with a transceiver

AF Stages

The AF output from an envelope or product detector of a superhet receiver is usually of the order of 0.5 to 1V, and many receivers incorporate relatively simple one- or two-stage audio amplifiers, typically using an IC device and providing about 2W output. On the other hand the direct-conversion receiver may require a high-gain audio section capable of dealing with signals of less than 1μV.

Provided that all stages of the receiver up to and including the product detector are substantially linear, many forms of post-

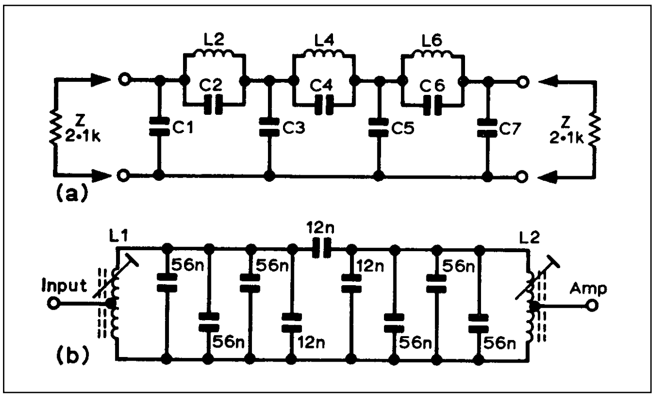


Fig 6.79: (a) Phone and (b) CW AF filters suitable for use in direct-conversion or other receivers requiring very sharply defined AF responses. The CW filter is tuned to about 875Hz. Values for (a) can be made from preferred values as follows: C1 37.26nF (33,000 + 2200 + 1800 + 220pF); C2 3.871nF (3300 + 560pF); C3 51.87nF (47000 + 4700 + 150pF); C4 19.06nF (18,000 + 1000pF); C5 46.41nF (39,000 + 6800 + 560pF); C6 13.53nF (12,000 + 1500pF); C7 29.85nF (27,000 + 2700 + 150pF). All capacitors mica or polyester or styroflex types. L2 168.2mH (540 turns), L4 124.5mH (460 turns); L6 129.5mH (470 turns) using P30/19 3H1 pot cores and 0.25mm enam wire. Design values based on 2000Ω impedance

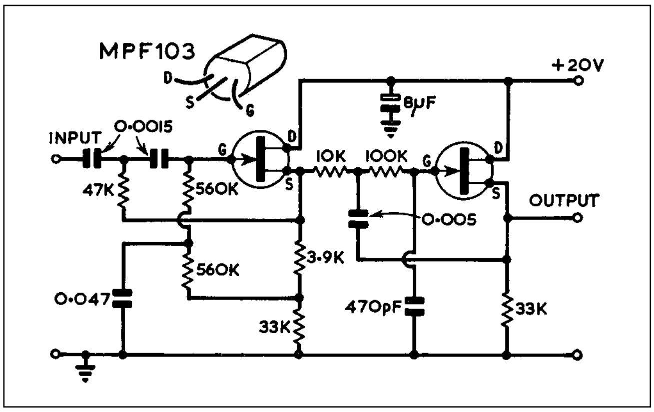


Fig 6.80: Active band-pass AF filter for amateur telephony. -6dB points are about 380 and 3200Hz, -18dB about 160 and 6000Hz

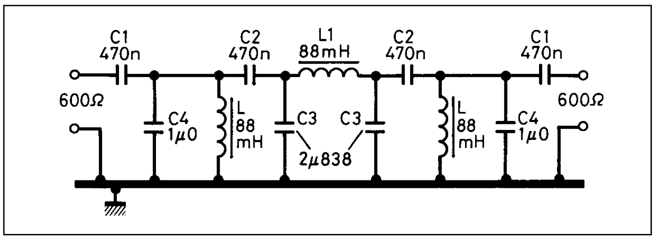


Fig 6.81: Passive AF filter by DJ1ZB using standard 88mH toroids and with a centre frequency of about 420Hz and bandwidth about 80Hz. Note that this design is for 600-ohm input/output impedance [*Sprat* No 58]

demodulation signal processing are possible: for example band-pass or narrow-band filtering to optimise signal-to-noise ratio of the desired signal, audio compression or expansion; the removal of audio peaks, AF noise blanking, or (for CW) the removal by gating of background noise. Audio phasing techniques may be used to convert a DSB receiver into an SSB receiver (as in two-phase or third-method SSB demodulation) or to insert nulls into the audio pass-band for the removal of heterodynes. Then again, in modern designs the AGC and S-meter circuits are usually operated from a low-level AF stage rather than the IF-derived techniques used in AM-type receivers.

It should be appreciated that linear low-distortion demodulation and AF stages are necessary if full advantage is to be taken of such signal processing, since strong intermodulation products can easily be produced in these stages. Thus, despite the restricted AF bandwidth of speech and CW communications, the intermodulation distortion characteristics of the entire audio section should preferably be designed to high-fidelity audio standards. Very sophisticated forms of audio filtering, notching and noise reduction are now marketed in the form of add-on digital-signal-processing (DSP) units, or may be incorporated into a transceiver.

Audio filters may be passive, using inductors and capacitors, or active, usually with resistors and capacitors in conjunction with op-amps or FETs (Figs 6.79-6.82). Many different circuits have been published covering AF filters of variable bandwidth, tuneable centre frequencies and for the insertion of notches. The full theoretical advantage of a narrow-band AF filter for CW reception may not always be achieved in operational use: this is because the human ear can itself provide a 'filter' bandwidth of about 50Hz with a remarkably large dynamic range and the ability to tune from 200 to 1000Hz without introducing 'ringing'.

MODIFICATIONS TO RECEIVERS

While the number of amateurs who build their own receivers from scratch is today in a minority, some newcomers or those with limited budgets buy relatively low-cost models or older second-hand receivers and then set about improving the performance. Old, but

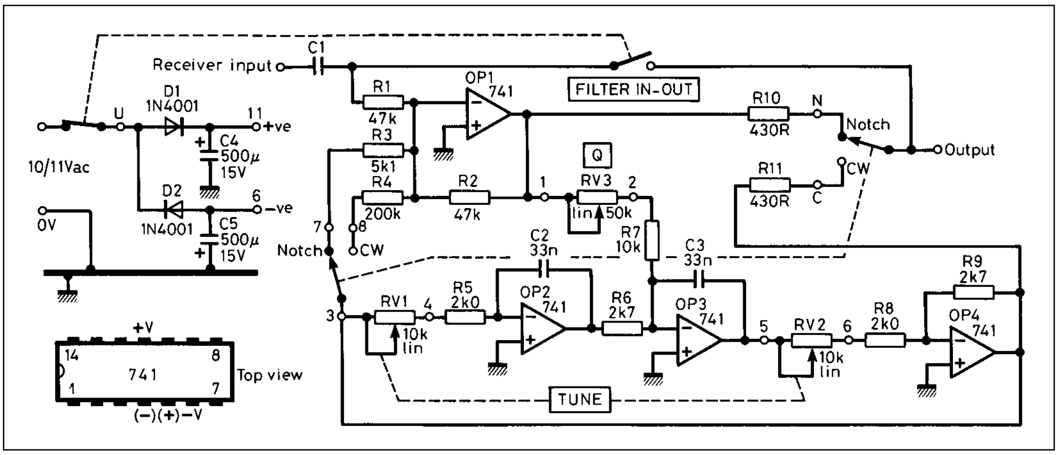


Fig 6.82: Versatile active analogue AF filter for speech or CW reception as described originally by DJ6HP in 1974 and which continues to represent an effective design. It provides a CW filter tuneable over about 450 to 2700Hz with the Q (bandwidth) variable over a range of about 5:1. For speech the filter can be switched to a notch mode. Although modern digital audio filters could provide more precisely shaped tuneable filtering, this analogue filter has received many endorsements over the years

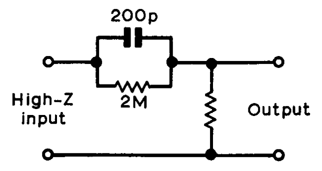


Fig 6.83: The single-crystal filter can be used effectively for phone reception by incorporating AF tone correction to remove the 'wooliness' of the heavily top-cut speech. A simple network such as the above provides top lift that restores intelligibility when used with the response curve of a typical single-crystal filter

basically well-designed and mechanically satisfactory, valved receivers can form the basis of excellent receivers, often rather better than is possible by modifying some more recent low-cost receivers. The main drawback of the older receivers is their long warm-up period, making it difficult to receive SSB signals satisfactorily until the receiver has been switched on for perhaps 15 or 20 minutes.

Some of the older models using relatively noisy mixer stages may be improved on 14, 21 and 28MHz by the addition of an external preamplifier and such a unit may also be useful in reducing image and other spurious responses. However, high-gain preamplifiers should not be used indiscriminately since on a low-noise receiver they will seriously degrade the signal-handling capabilities without providing a worthwhile improvement of signal-to-noise ratio.

Receivers having low noise but poor signal-handling capabilities can more often be improved by the fitting of a switched, adjustable or AGC-controlled antenna attenuator. Such an attenuator is likely to prove of most use on 7MHz where the presence of extremely strong broadcast signals will often result in severe cross-modulation and intermodulation.

A receiver deficient in selectivity can often be improved by adding a second frequency changer followed by a low-frequency (50 to 100kHz) IF amplifier (a technique sometimes known as a Q5-er); or by adding a crystal or mechanical filter, or by fitting a Q-multiplier. CW reception can be improved by the use of narrow-band audio filters, although the degree of improvement may not always be as much as might be expected theoretically because of the ability of an experienced operator to provide a high degree of discrimination.

Older receivers having only envelope detection may be improved for SSB and CW operation by the fitting of a product detector and possibly adding a good mechanical or crystal band-pass filter. NBFM reception may be included by adding an FM discriminator.

Many older receivers use single rather than band-pass crystal filters (and the excellence of the single crystal plus phasing control for CW reception should not be underestimated) and these often provide a degree of nose selectivity too sharp for satisfactory AM or SSB phone reception: speech may sound 'woolly' and virtually unintelligible due to the loss of high- and low-frequency components. However, because the response curve of such filters is by no means vertical, the addition of a high degree of tone correction (about 6dB/octave) can do much to restore intelligibility and the combination then provides an effective selectivity filter for SSB reception. The tone correction circuit shown in **Fig 6.83** is suitable for high-impedance circuits and can be adapted by using higher C and lower R for low-impedance circuits.

The addition of an antenna matching unit between receiver and antenna can improve reception significantly in those cases where appreciable mismatch may exist (for example when using random-length long-wire antennas with receivers intended for use with a 50 or 70Ω dipole feeder) (**Fig 6.84**).

A common fault with older receivers is deterioration of the Yaxley-type wave-change switch and/or the connection to the rotor spindle of the variable tuning capacitors; such faults may often cause bad frequency instability and poor reset performance. Improvement is often possible by the careful use of modern switch-cleaning lubricants and aerosols.

A simple accessory for older receivers (or those modern receivers not already incorporating one) is a crystal calibrator providing 'marker' signals derived from a 100kHz or 1MHz crystal. While a simple 100kHz oscillator will usually provide harmonics throughout the HF range, the availability of integrated circuit dividers makes it practicable to provide markers which are not direct harmonics of the crystal. For example 10kHz or 25kHz or even 1kHz markers can be provided using TTL decade divider logic or divide-by-two devices.

A receiver deficient in HF oscillator stability on the higher frequency bands may still form the basis of a good tuneable IF strip when used on a low frequency band in conjunction with a crystal-controlled converter. Again, when the basic problem is oscillator drift due to heat, this can sometimes be reduced by fitting silicon power diodes in place of a hot-running rectifier valve or by adding temperature compensation to the HFO. A more drastic modification is to replace an existing valve HFO with an internal or external FET VFO. Excessive tuning rate can sometimes be overcome by fitting an additional or improved slow-motion drive. A receiver with a good VFO can be modified for really high-stability performance (better than about 20Hz) by means of external 'huff and puff' digital stabilisation using crystal-derived timing periods. See the chapter on oscillators in this book and [14, 15].

In brief, the excellent mechanical and some of the electrical characteristics of the large and solidly built receivers, such as the AR88, HRO, SUPER-PRO and some Eddystone models which featured single conversion with two tuned RF stages, are seldom equalled in modern 'cost-effective' designs. It may prove well worth spending time and trouble to up-grade these vintage models into receivers which can be excellent even by modern standards. Post-war Collins and Racal RA17L valved receivers remain highly regarded HF receivers.

The following summary indicates some common faults with older models and ways in which these can be overcome.

Poor sensitivity

Due to atmospheric noise this usually only degrades performance on 21 and 28MHz and then only on older valve models. Sensitivity can be improved by the addition of a preamplifier, but gain should not be more than is necessary to overcome receiver noise. Note that the sensitivity of a receiver may have been impaired by poor alignment, or by mismatched antennas, or due to the ageing of valves.

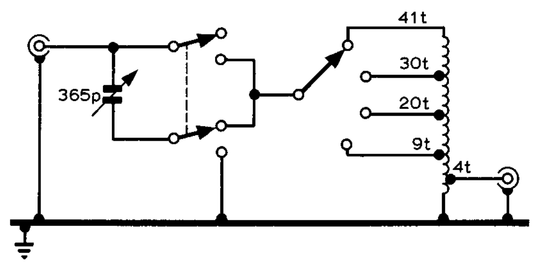


Fig 6.84: Typical 0.55 - 30MHz antenna tuning and matching unit

Image response

This can be reduced by additional pre-mixer selectivity, often most conveniently by means of a low-gain preamplifier with two or more tuned circuits. It is also possible to use a pre-tuned filter such as the Cohn minimum-loss filter for particular bands.

Stability

This is a direct function of the oscillators within the receiver. Excessive drift and frequency 'jumping' may be due to a faulty valve or band-change switch, or to incorrect adjustment of any temperature-compensation adjustments. Drift can sometimes be reduced by reducing the amount of heating of the oscillator coil by fitting heat screens, or by the addition of temperature compensation.

But often with older receivers it will be found difficult to achieve sufficient stability on the higher frequency bands. In such cases considerably greater stability may be achieved by using the receiver as a variable IF system on one of the lower frequency bands, with the addition of one or more crystal-controlled converters for the higher frequency bands. It is worth noting that all oscillators (not only the first 'HFO') may be the cause of instability (eg second or third frequency conversion oscillator or even the beat frequency oscillator).

ZS5JF cured slow drift on a hybrid radio by removing the internal loudspeaker and installing a 12v computer fan in its place to extract the warm air from around a valve.

Tuning rate

The tuning rate of some older but still good receivers tends to be too fast for easy tuning of SSB and CW signals. Often this problem can be overcome by the fitting of an additional slow-motion drive on the main tuning control.

Selectivity

It is possible to improve the selectivity of a receiver by fitting an external low IF section, or by fitting a (better) crystal filter, or a Q-multiplier. Many SSB receivers make little provision for narrow-band CW reception and in such cases it may be possible to include a single-crystal filter with phasing control in one of the later IF stages, or to add a Q-multiplier or audio filter.

Blocking and intermodulation

Performance of many semiconductor (and some valve) receivers can be improved by the addition of even a simple antenna attenuator for use on 7MHz in the presence of extremely strong broadcast signals.

MEASUREMENTS AND TEST EQUIPMENT

Generally speaking, the complexity of the project is governed by the available test equipment. The essentials are a multimeter, a power supply, an RF signal generator, an audio generator, a frequency counter and an oscilloscope. The 'nice to have' equipment includes a spectrum analyser with tracking generator, a Q-meter, a variable attenuator and a second RF signal generator. Such equipment is readily available second hand, to be found at rallies, internet auction sites and specialist dealers [16].

Direct conversion receivers can be built successfully using the above essential test equipment whilst it is difficult to design or align superhets without at least a spectrum analyser and its tracking generator.

Designing inductors is made much easier with the aide of a Q-meter. These are harder to find second hand but there have been a few designs to make your own published in the magazines [17]. Alternatively the dip oscillator can be used [18].

Spectrum analysers have always commanded a high price due to their complexity and feature list. For HF use, quite modest analysers are sufficient such as the HP-141T family with its separate tracking generator and the Marconi TF2370 with a built in tracking generator.

Much more on this subject can be found in the chapter on Measurement and Test Equipment

BUILDING RECEIVERS

For many years the percentage of home-built HF receivers in use on the amateur bands has been very low and increasingly has been confined to specialised sectors of the hobby such as compact, low-power (QRP) portable operation, often on one specific band and for a single mode of operation (often CW only) or as introductory receivers for those to whom the rather daunting cost of factory-built receivers or transceivers can represent a deterrence to HF operation. There remain, however, valid reasons to encourage home-construction, not only for those with limited budgets but also as an ideal form of 'hands-on' learning process.

A major advantage of building or modifying older receivers is that the constructor can then be confident of maintaining it in good trim. It is a substantial advantage for the amateur, particularly if located a long way away from the suppliers, to use equipment that he or she feels capable of keeping in good condition, and carrying out his or her own repairs when necessary. Increasingly factory-built equipment for the amateur does not lend itself to home-servicing.

Nor should it be forgotten that for those with the necessary practical experience, the availability of complex integrated circuits developed for consumer electronics and ceramic resonators that do not require skilled alignment has eased the construction of both simple and high-performance receivers. The factory designer must usually cater for all possible modes and bands, usually said to contain all the 'bells and whistles', most of which are not used in practice, whereas the constructor can build a no-compromise receiver to suit precisely his or her own particular interests.

The amateur can still provide himself with a station receiver or transceiver, or a portable receiver or transceiver that can bear comparison with the best available factory models, and in doing so prove that the communications receiver is not a 'black box' or 'consumer appliance' of which the technology remains a largely unknown quantity.

It has traditionally been a feature of the hobby of amateur radio that the enthusiast strives to understand the technology; the home-construction and home-maintenance of equipment are not assets that should be surrendered lightly despite the undoubted attractions of factory-designed equipment.

Simple receivers, converters and single-band QRP transceivers can be built in an evening or two, but an advanced receiver - or receiver section of a transceiver - may take several months of work and adjustment.

Although the homodyne-type direct conversion receiver and the 'zero-IF' form in which the output from a first mixer is fed directly into a product detector make possible low-cost HF receivers which are simple to build, care must be taken if optimum performance is to be achieved.

Because the necessary high overall gain (about 100dB) is achieved virtually entirely in the AF amplifier, these stages are very sensitive to hum pick-up at the 50Hz mains supply frequency and its harmonics. The AF stages should be well decoupled with the mixer and first AF stage in close proximity to enable the connections between them to be as short as possible; power supplies need to be well smoothed with care taken

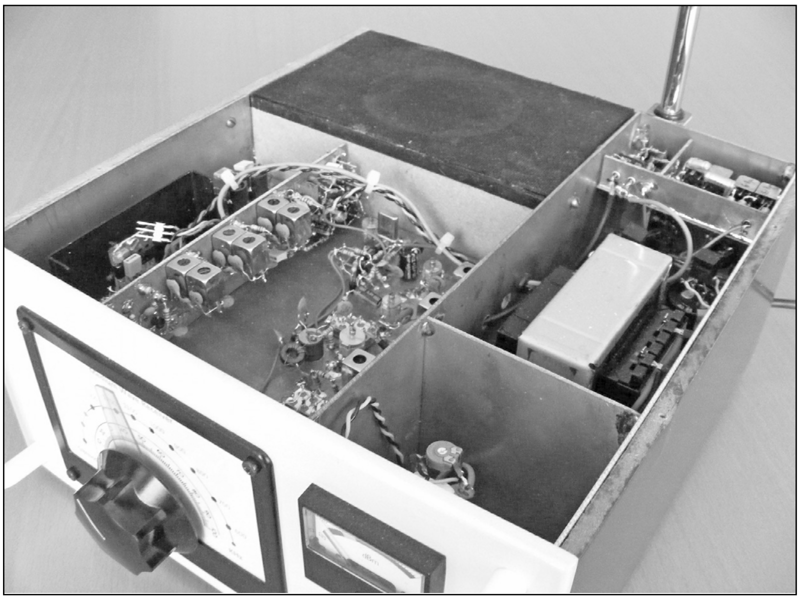


Fig 6.85: PCB used as a structural material

not to introduce 'earth loops' and/or direct pick-up from the magnetic field surrounding mains transformers and inductors. AF hum can usually be avoided altogether with battery-operated receivers.

The high AF gain may also introduce 'microphony' with components acting as microphones. A loudspeaker tends to make components vibrate and headphones are preferable, with the added advantage of requiring less AF output and overall gain. Ceramic capacitors, which exhibit piezo-electric characteristics, tend to introduce microphony; moulded polycarbonate capacitors are much to be preferred. Similarly, ferrite toroid cores used for AF filters can also introduce microphony. For inductors with values greater than 0.1mH, screened air-cored inductors are preferable.

Direct-conversion receivers of all types, including those with regenerative detectors, tend to suffer from local oscillator radiation from the receiving antenna and can cause interference in the locality, unless RF leakage through even a double-balanced mixer/demodulator is reduced, eg by using a broadband isolator. A low-gain resonant or broad-band RF stage reduces radiation.

Microphony may also arise in the signal-frequency components, with the mixer acting as a phase detector, reflecting the interaction between the LO radiation or leakage and the incoming signal in a high-Q tuned circuit. Remember that the LO voltage is very much stronger than the incoming signal and it may be difficult to eliminate RF microphony altogether without effective screening and isolation.

Oscillator radiation can also result in RF hum that appears as 50Hz hum in the headphones. This arises from the oscillator signal becoming hum modulated in the mains wiring/rectifiers and then being re-radiated back to the receiver antenna. The cure is to stop the local oscillator signal from radiating; all connections to the receiver, such as power supply and headphones, should be RF grounded to the receiver case using decoupling capacitors.

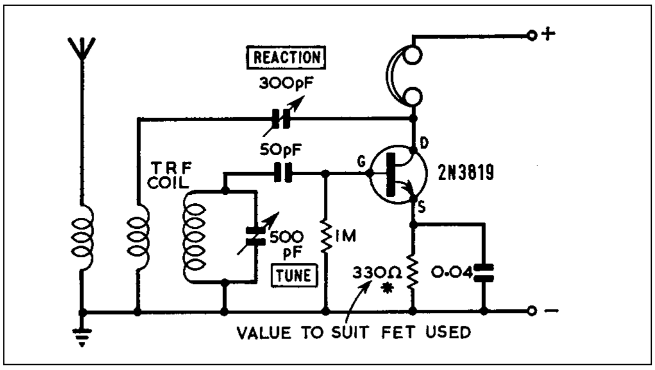


Fig 6.86: Single-FET receiver

Using PCB as a Structural Material

Printed circuit board, particularly the glass fibre based board makes an excellent chassis type material, combining both strength and solderability. It is easy to cut, drill and shape although a face mask is recommended regarding the resulting dust. It lends itself to the 'dead bug' style of construction where components are soldered directly to each other or to ground, see **Fig 6.85** and the chapter on Construction and Workshop Practice.

Designs for Home Construction

The simplest types of HF receiver suitable for home construction 'on the kitchen table' without the need for high-grade measuring equipment etc are undoubtedly the various forms of 'straight' (direct-conversion) receivers using either a regenerative detector, or Q-multiplier plus source-follower detector, or the now more popular homodyne form of direct conversion, preferably using a balanced or double-balanced product detector. In practice, to avoid the complications of band-switching, most of the simple models tend to be either single-band receivers or may still use the once-popular 'plug-in' coils.

Figs 6.86-6.91 show a representative selection of circuit diagrams for simple receivers, suitable for use directly with high-impedance headphones, or with output transformers permitting the use of modern low-impedance headphones. As mentioned in the previous section, simple regenerative receivers of the type shown in Figs 6.86 and 6.87 can re-radiate the local oscillator signal

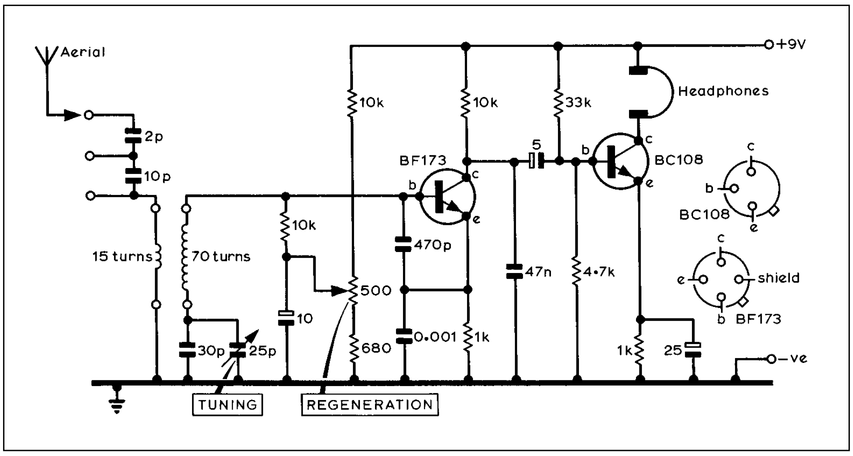


Fig 6.87: A simple 'straight' receiver intended for 3-5MHz SSB/CW reception and using Clapp-type oscillator to improve stability

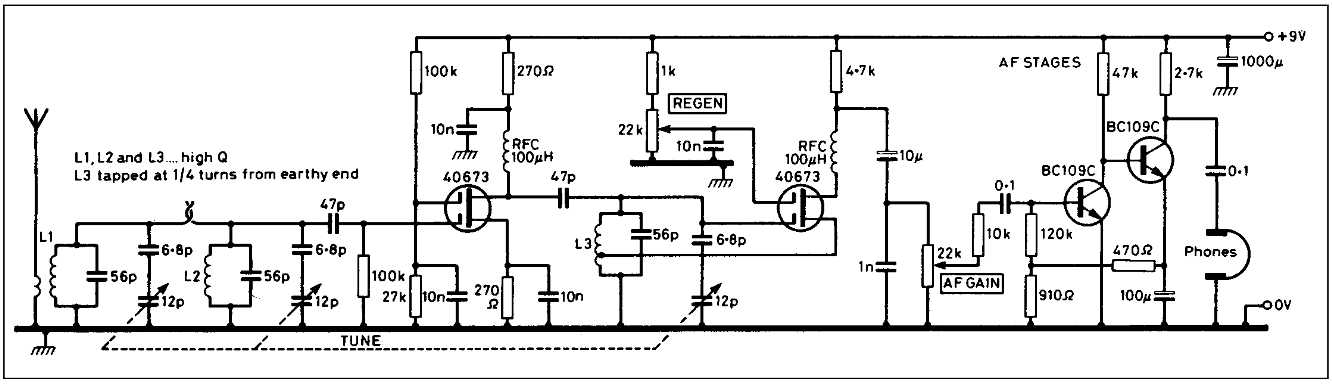


Fig 6.88: Solid-state 14MHz TRF 'straight' receiver originally described by F9GY in *Radio-REF* in the 1970s and intended primarily as a monoband CW receiver

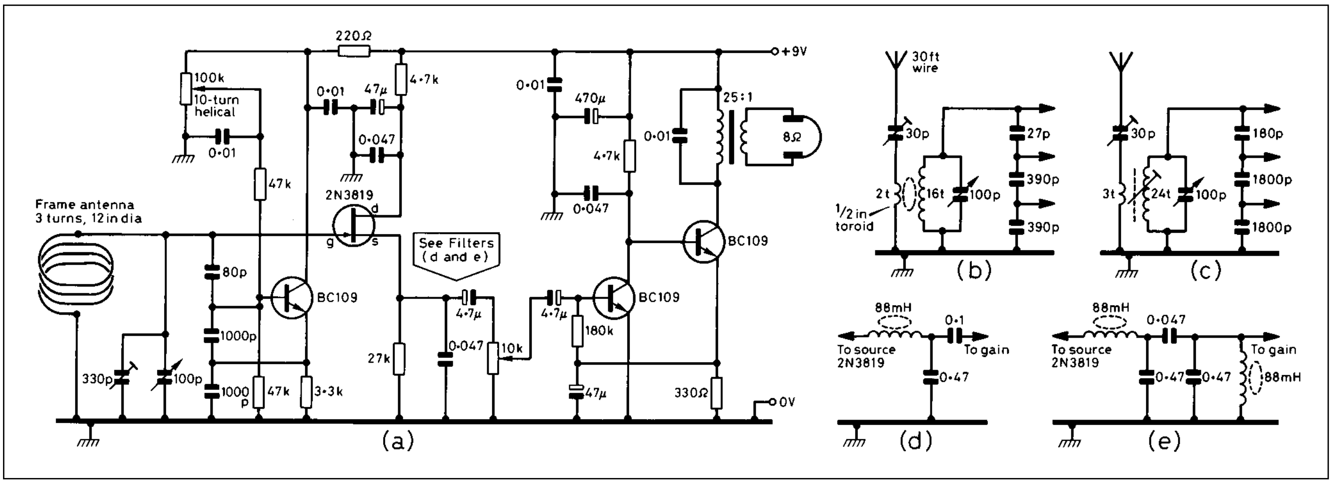


Fig 6.89: GI3XZM's solid-state regenerative 'blooper' receivers. (a) 3.5MHz version with frame antenna mounted about 12in above chassis with miniature coaxial-cable 'downlead'. (b) Input circuit for 9-16MHz version. (c) Conventional input circuit for 3.5MHz receiver using wire antenna (coil 26SWG close-wound on 0.5in slug-tuned former). (d) Audio filter that replaces the 4.7μF capacitor shown in (a). (e) CW filter 14MHz Mk2 version

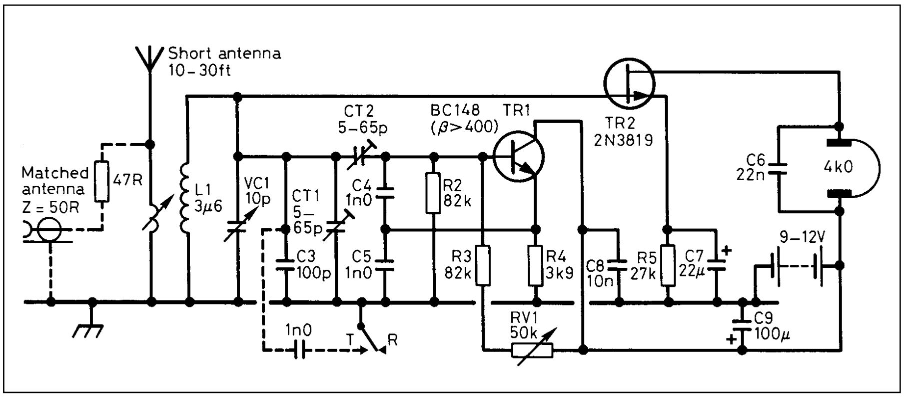


Fig 6.90: G3RJT's 'two-transistor communications receiver' or 'active crystal-set receiver', based on the design approach of GI3XZM but using a higher-gain drain-bend detector that permits the omission of the two-transistor AF amplifier provided that high-impedance headphones of good sensitivity are used

which can cause interference over a sometimes considerable area. Care may need to be taken to ensure this does not occur.

For simple receivers where a very high dynamic range is not sought, construction can be simplified by the use of SA602 type IC devices which contain a double-balanced mixer, oscillator and isolator stages. This device, originally developed for VHF portable radiophones, has been widely adopted by amateurs for use as frequency converters, complete front-ends for direct-conversion receivers, and as product detectors etc.

Fig 6.92 outlines the SA602 and some ways it can be used, while **Fig 6.93** shows its use as the front-end of a 28MHz direct-conversion receiver which could be adapted for lower HF bands.

The SA602 is equally suitable for use as a crystal-controlled frequency converter to provide extra bands in front of an existing receiver or for the 'super-gainer' form of simple superhet or as the mixer/oscillator stages of a conventional superhet, possibly using a second SA602 as a product-detector.

Receivers based on standard IC devices have particular application for 'listening' or as the receiver section of compact low-power transceivers.

Fig 6.94 shows how a Motorola MC3362 IC can form the complete front-end of a single-band superhet including IF and detector stages as part of a compact transceiver. The chip pinout is in **Fig 6.95**.

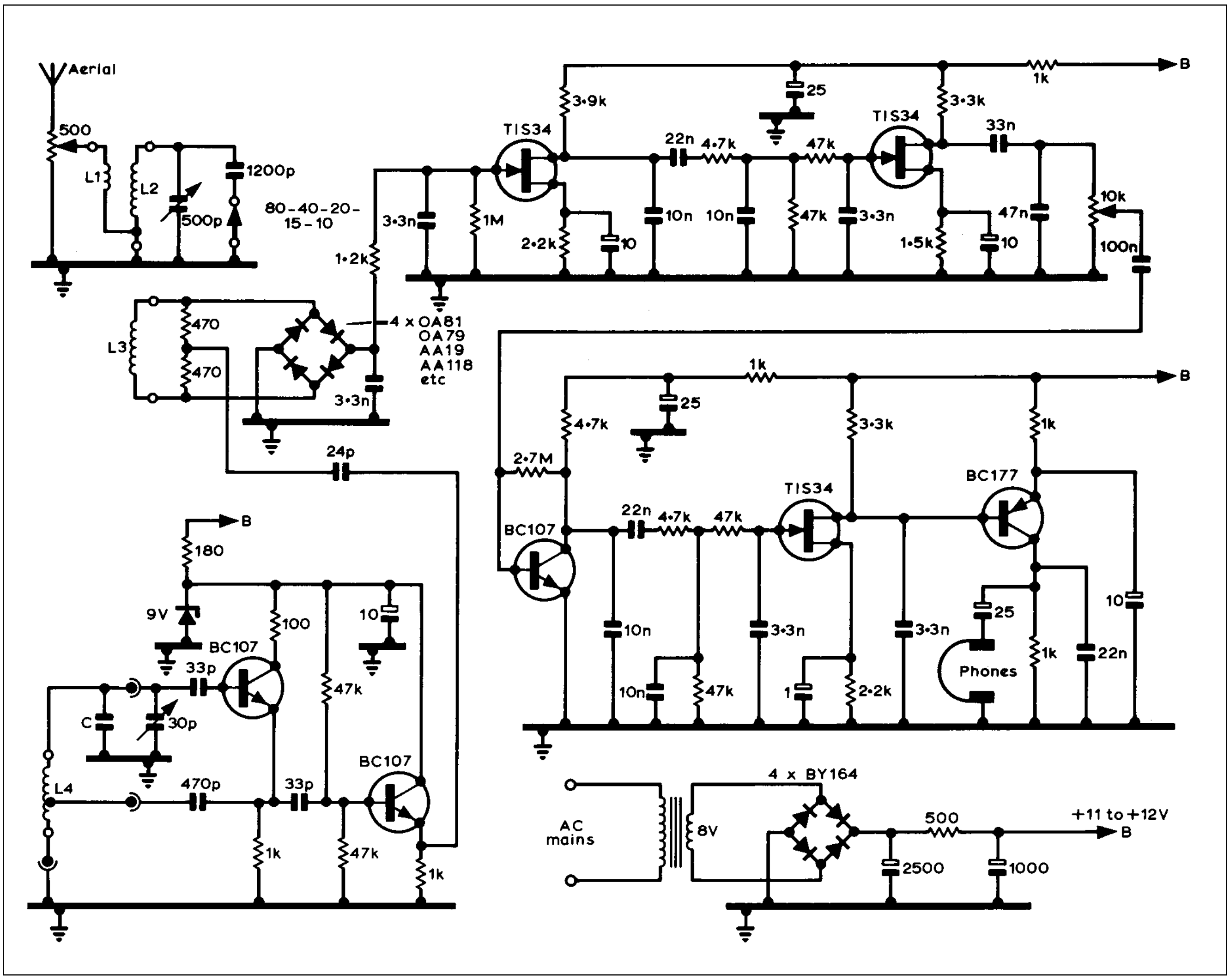


Fig 6.91: Complete circuit diagram of a multiband direct-conversion receiver using diode ring demodulator and plug-in coils for oscillator section

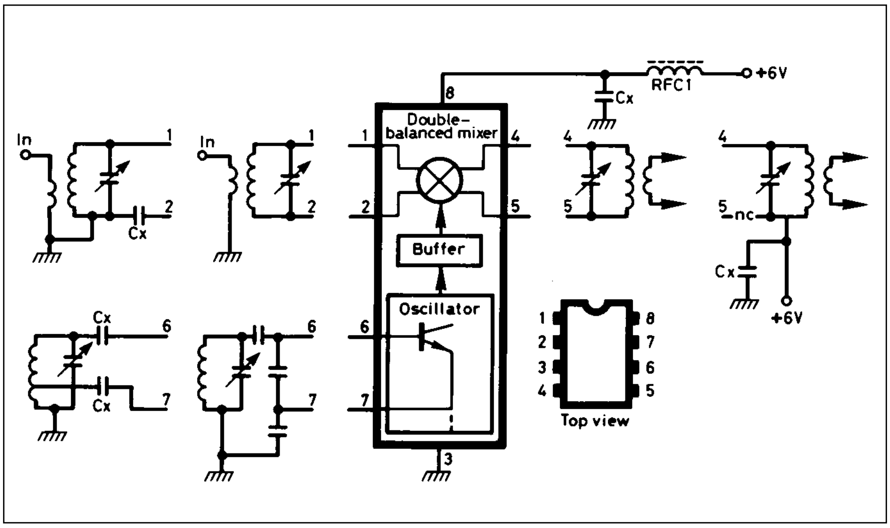


Fig 6.92: Typical configurations of the SA602. Balanced circuits are to be preferred but may be more difficult to implement. Cx blocking capacitor 0.001 to 0.1μF depending on frequency. RFC1 (ferrite beads or RF choke) recommended at higher frequencies. Supply voltage should not exceed 6V (2.5mA). Noise figure about 5dB. Mixer gain 20dB. Third-order intercept +15dBm (do not use an RF preamplifier stage). Input and output impedances are both 2 x 1.5kΩ.

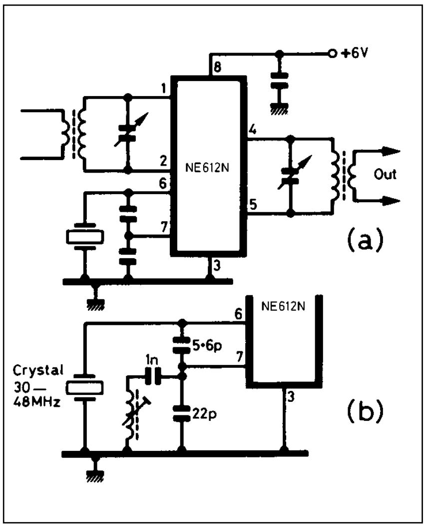


Fig 6.93: The use of an SA602 IC as a crystal-controlled converter. The 5dB noise figure is low enough to achieve optimum sensitivity right up to about 50MHz without having to build an RF amplifier

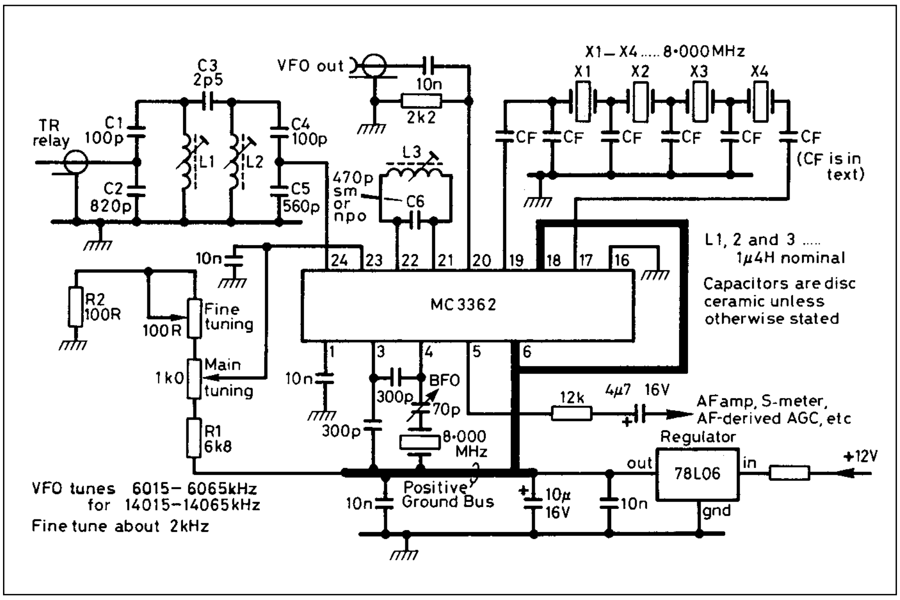


Fig 6.94: K9AY uses the MC3362 as the complete front-end of the 14MHz superhet receiver

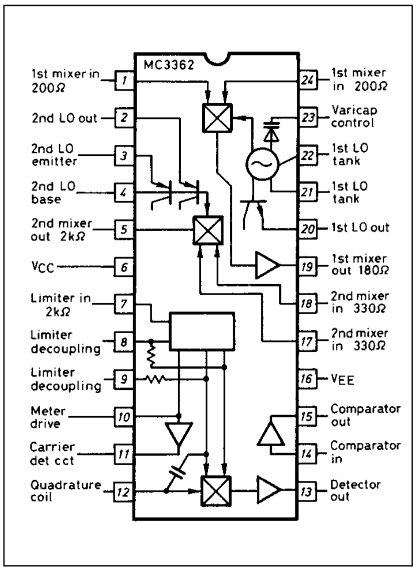


Fig 6.95: The MC3362 chip showing pin-out and basic functions

Super-linear Front-ends

The front-end of a superhet or direct-conversion receiver comprises all circuitry preceding the main selectivity filter. For a superhet this includes the passive preselector, the RF amplifier(s), the mixer(s) and heterodyne oscillator(s), the diplexer between mixer and post-mixer amplifier, the roofing filter, and any IF stages up to an including the main (crystal) filter. For a direct-conversion receiver, the front-end comprises all stages up to the selective audio-filter(s), including the product detector (which for a high-performance receiver may be of the two-phase, audio-image-rejecting type).

For any receiver, superhet or direct-conversion, in which digital signal processing at IF or audio baseband is used to determine the selectivity, the A/D converter and digital filtering must be considered in determining the front-end performance in terms of linearity and dynamic range.

In designing a receiver for the highest possible front-end performance, attention must be paid to all of the circuitry involved, to the gain distribution, to the noise characteristics, to both the strong-signal handling characteristics and to the intermodulation intercept points.

The ability to hold and copy an extremely weak signal, barely above the atmospheric noise level, adjacent to a strong local signal or close to signals from super-high-power broadcast signals places heavy demands on the active and passive components available within amateur budgets, including any ferrite-cored transformers and the crystal filters.

The limiting factors in the design of high-performance, solid-state HF receivers remain the spurious-free dynamic range (SFDR) of the mixer-stage, the noise and stability of the associated oscillator and strong-signal performance of the filters. Jacob Makhinson, N6NWP [19] believes, by applying known design principles, radio amateurs can construct a high-performance front-end which combines a very high intercept point with excellent sensitivity.

Used with a low-noise local oscillator, a front-end based on a DMOS FET quad device as a double-balanced switching mixer and low-noise square-wave injection, obtained by means of a dual flip-flop followed by a simple diplexer network at an IF of 9MHz, can achieve a wide dynamic range even when a suitable push-pull RF low-noise amplifier is used ahead of the mixer. N6NWP stresses that: "A receiver incorporating such a front-end can provide strong-signal performance that rivals or exceeds that of most commercial equipment available to the amateur."

Investigations of the N6NWP front-end (Fig 6.96) have found that it is possible to achieve extremely high third-order intercept points (+50dBm on 1.8, 3.5 and 7MHz, +45dBm on 14MHz subject to some spread in devices in different batches) to the extent where other parts of the circuitry, such as the diplexer or crystal filter rather than the mixer, tend to become limiting factors.

The SD8901 device (available from Calogic Corporation) contains four DMOS FETs configured for use as a commutation (switching) mixer (see Figs 6.60-6.62).

N6NWP utilises this device with square-wave drive to the gates of the FETs from a high-speed CMOS D-type bistable

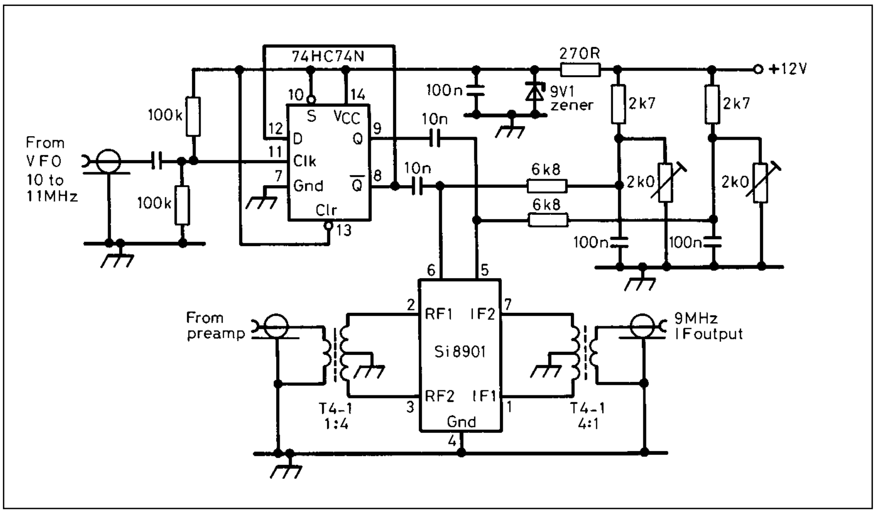


Fig 6.96: The basic high-dynamic-range MF/HF receiver front-end mixer circuitry as developed by N6BWP

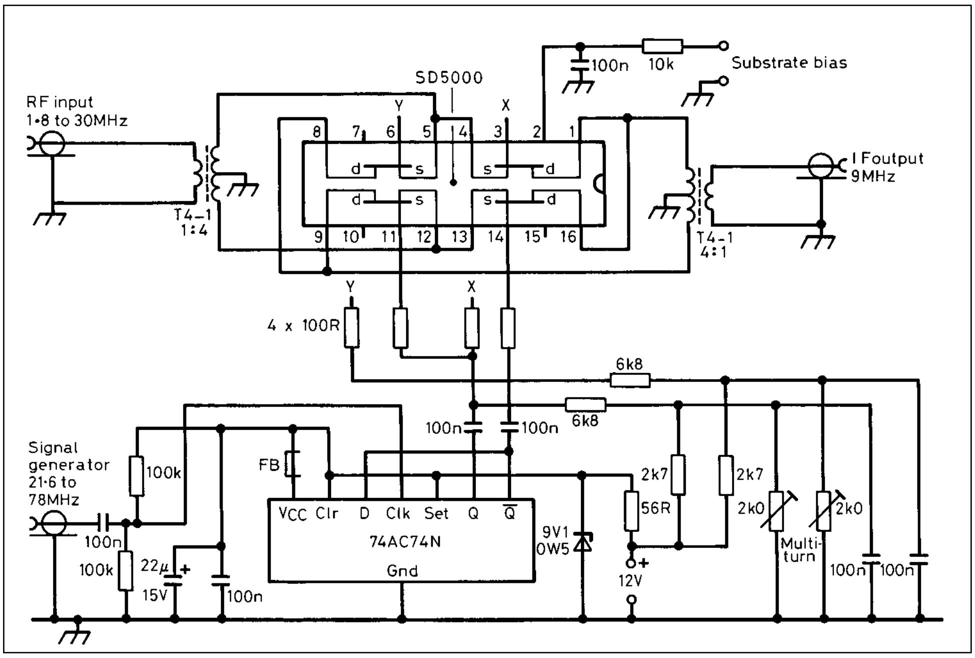


Fig 6.97: G3SBI's modified N6NWP-type mixer test assembly using the SD5000 FET array and 74AC74N to provide square-wave injection from a high-quality signal generator source at twice the required frequency

device operating unusually from a 9V supply. The 14MHz intercept point is about +39dBm, an excellent figure.

The initial mixer investigated by G3SBI was based on the N6NWP approach but used the more widely available Siliconix SD5000 quad DMOS FET array (batch 9042). Since this array has gate-protection diodes, the substrate needs to be biased negatively to prevent gate conduction under some conditions; however, the array has the advantage of close matching of the drain-to-source on-resistance.

The test board using the circuit of Fig 6.97 was made using earth-plane construction, with all transformers and ICs fitted into turned-pin DIL construction, so that they could be changed easily. With the test set-up of Fig 6.98, it initially proved possible to achieve a true input intercept of +42dBm on 14MHz using 5MHz local oscillator injection. An input intercept of +45dBm was obtained on 3.5MHz with a 5.5MHz local oscillator frequency. With a local oscillator running at 23MHz the 14MHz intercept was a few dBm down compared with the 5MHz local oscillator. For this reason an advanced CMOS 74AC74 device was used as the LO squarer, resulting in near-perfect 50-50 square waveform.

To reduce ringing, only one D-type in the chip was used and stopper resistors were connected to the FET gates; a single ferrite bead in series with the Vcc pin proved useful. It is important for the oscillator injection to be a clean square wave if the results given above are to be obtained. With these modifications

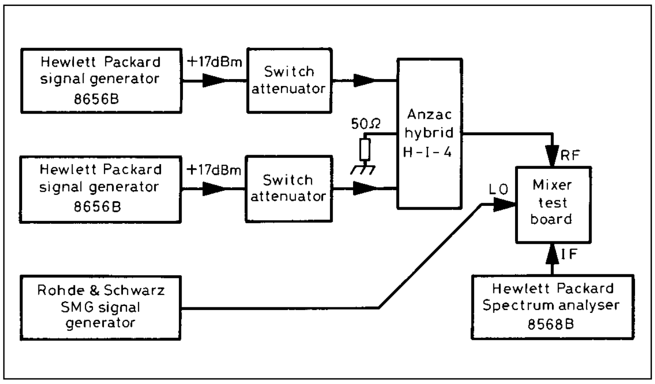


Fig 6.98: The test instrumentation used by G3SBI for intermodulation tests on the mixer

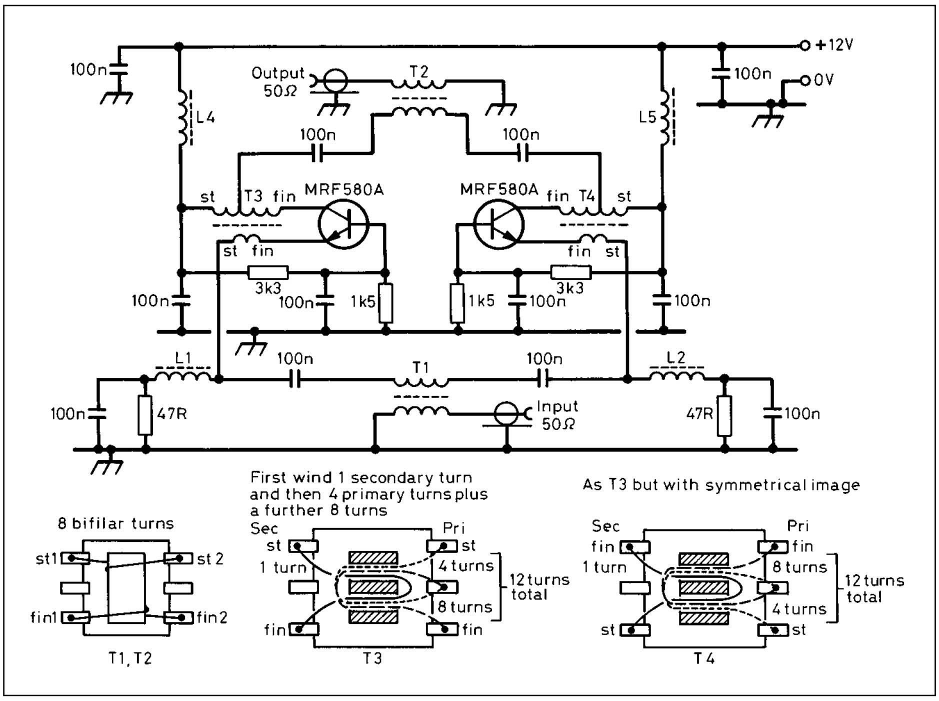


Fig 6.99: G3SBI's modified post-mixer test amplifier adapted from the N6NWP design but using the MRF580A devices. All resistors 0.25W metal-film RS Components. All 0.1µF capacitors monolithic ceramic RS Components. L4, L5 4t of 0.315mm dia bicelflux wire on RS Components ferrite bead. L1, L2 5t 0.315mm dia bicelflux wire (RS Components). T1-T4 use 40swg bicelflux wire. Take two glassfibre Cambion 14-pin DIP component headers, cut each into two parts and bend the tags 90° outwards. Stick a piece of double-sided tape onto the header and mount the balun cores on this. Wind the transformers as shown above. The amplifier is constructed with earth-plane layout

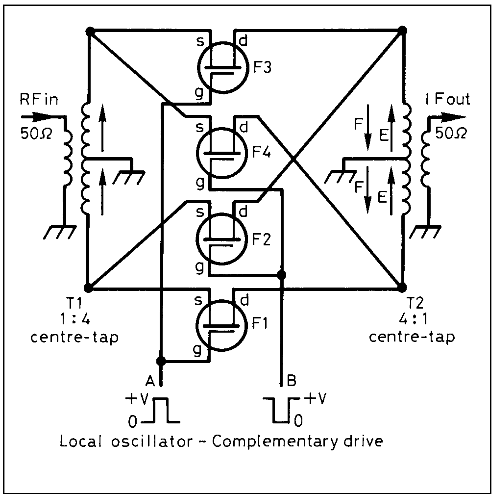


Fig 6.100: Conventional commutation ring mixer

input intercepts of at least +42dBm were achieved on all HF bands and +46dBm on 1.8 and 3.5MHz. On 7MHz no IMD was visible on the spectrum analyser, even with a bandwidth of 10Hz, representing an input intercept of +50dBm. A substrate bias of -7.5V and a gate bias of about +4.5V were used; conversion loss was 7dB. However, the intercept point was found to degrade sharply as soon as the input signal exceeds +7dBm (0.5V); the situation can be recovered by dropping the gate bias voltage, but it is then no longer possible to achieve intercept points above +45dBm.

Fig 6.99 shows a 9MHz post-mixer amplifier developed by G3SBI, again based on the N6NWP approach but with changes that provide improved performance in terms of gain and output intercept point, with a noise figure of 0.5dB. Whereas N6NWP used the MRF586 device, G3SBI used the MRF580A device, giving a lower noise figure at a collector current of 60mA. Measured performance showed a gain of 8.8dB, output intercept +56dBm, noise figure 0.5dB. However, a crystal filter driven by the amplifier would present a complex impedance, particularly on the slope and near the stop-band, and would seriously degrade performance of the amplifier.

G3SBI also investigated the performance of quadrature hybrid 9MHz crystal-filter combinations, and he found that the performance of budget-priced crystal lattice-type filters is a serious limitation (home-made ladder filters appear to have higher intercept points and lower insertion loss although the shape factor may not be quite so good).

The problem with budget-priced lattice filters can be reduced by eliminating the post-mixer amplifier with the mixer going immediately to a quadrature hybrid network 2.4kHz-bandwidth filter, followed by a low-noise amplifier. The 2.4kHz filter is then used as a roofing filter. Although this is not an ideal arrangement, it can result in an overall noise figure of about 13.5dB (5dB noise figure due to the filter and amplifier, another 7dB from mixer loss, and a further 2.5dB loss due to the antenna input band-pass filter). This would be adequate sensitivity on 7MHz without a pre-mixer amplifier.

The N6NWP-type mixer followed basically accepted practice in commutation (switching mixers) achieving a +50dBm input intercept point on 7MHz when used with a precise square-wave drive, but the performance fell off on bands lower and higher in frequency than 7MHz. Results could be improved on the lower frequency bands by altering the capacitive balance of the RF input and above, but this had no significant effect on 14MHz and above.

Fig 6.100 illustrates a conventional commutation ring mixer. If A is 'on', FETs F1 and F3 are 'on' and the direction of the RF signal across transformer T2 is given by the 'F' arrows. A deficiency of this arrangement is that as the RF input signal level increases, it has a significant effect on the true gate-to-source voltage needed to switch the FET 'on' or keep it switched 'off'. Larger local oscillator amplitudes are then required, but linearity problems may still exist because of the difference in the FET 'on' resistance between negative and positive RF signal states.

The new mixer developed by G3SBI (intellectual title held by SERC) is shown in **Fig 6.101** which illustrates why it has been given the name 'H-mode'. Operation is as follows: Inputs A and B are complementary square-wave inputs derived from the sine-

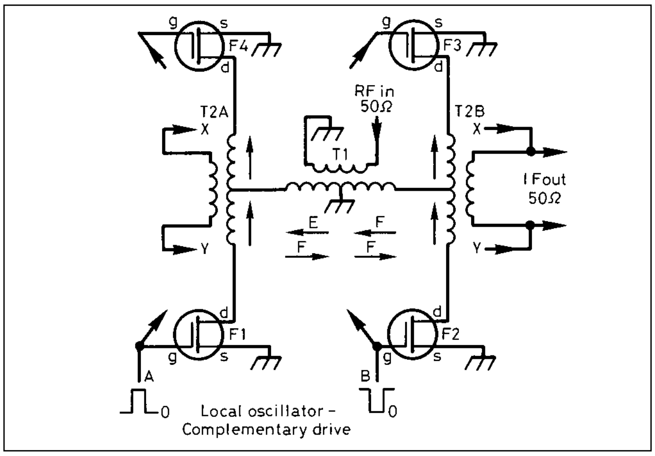


Fig 6.101: G3SBI 'H-mode' mixer

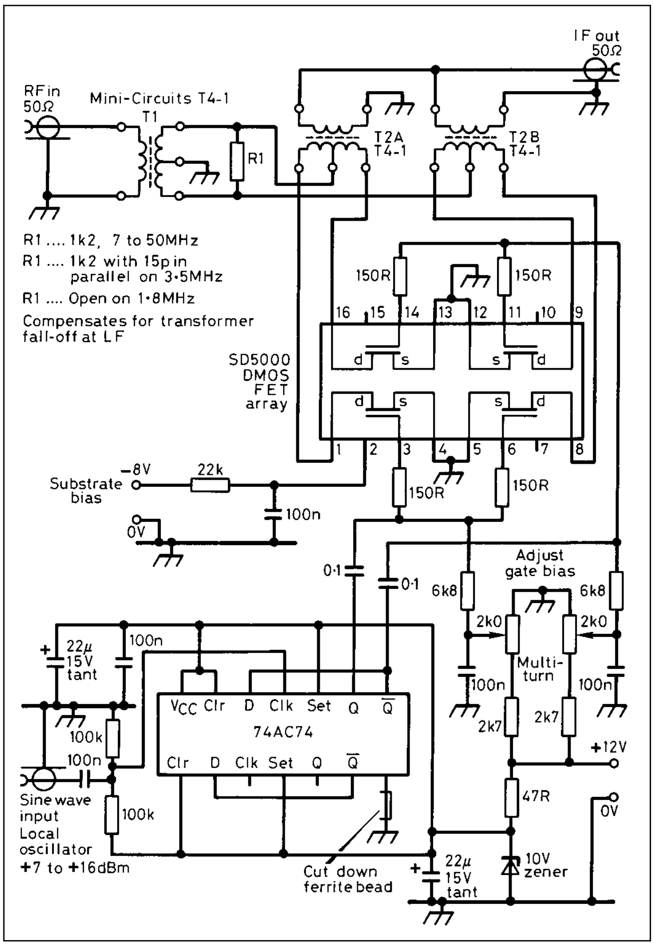


Fig 6.102: Test assembly for 'H-mode' mixer

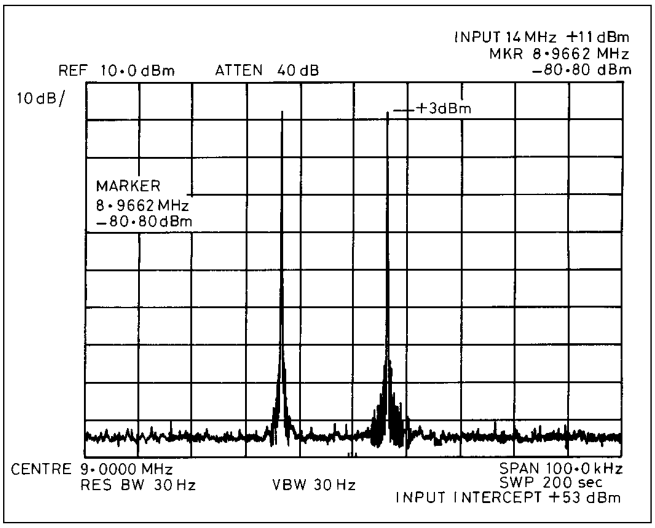


Fig 6.103: 14MHz input intermodulation spectrum

wave local oscillator at twice the required frequency. If A is 'on' then FETs F1 and F3 are 'on' and the direction of the RF signal across T1 is given by the 'E' arrows. When B is 'on', FETs F2 and F4 are 'on' and the direction of the RF signal across T1 reverses (arrows 'F'). This is still the action of a commutation mixer, but now the source of each FET switch is grounded, so that the RF signal switched by the FET cannot modulate the gate source voltage.

In this configuration the transformers are important: T1 is a Mini-Circuits type T4-1; T2 is two Mini-Circuits T4-1 transformers with their primaries connected in parallel. The parallel-connected transformers give good balance and perform well.

A practical test circuit of the H-mode mixer is shown in **Fig 6.102**. It was constructed on an earth plane board with all transformers and ICs mounted in turned-pin DIL sockets. The printed circuit tracks connecting T1 to T2 and from T2 to the SD5000 are kept short and of 0.015in width to minimise capacitance to ground.

The local oscillator is divided by two in frequency and squared by a 74AC74 advanced CMOS bistable similar to that used in the N6NWP-type mixer. However, the bistable is run from +10V instead of +9V and a cut-down RS Components ferrite bead is inserted over the ground pin of the 74AC74 to clean up the square wave.

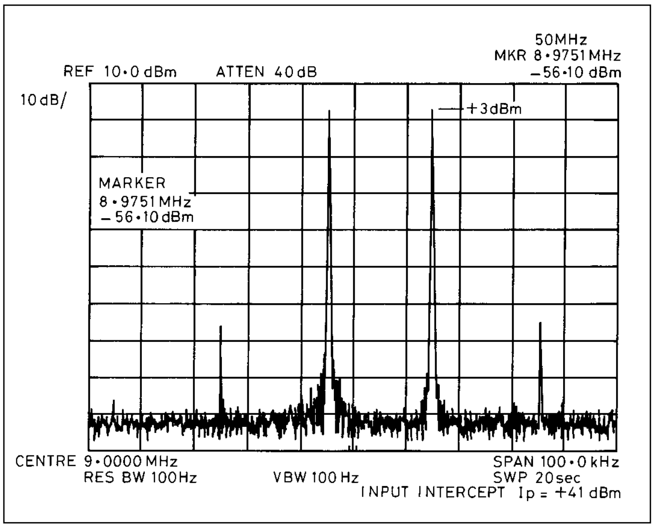


Fig 6.104: 50MHz input, output spectrum at 9MHz

The preferred method of setting the gate-bias potentiometers with the aid of professional-standard test equipment is as follows. One potentiometer is set to the desired bias voltage for a specific test run, the other is then set by looking at the RF-to-IF path feedthrough on the spectrum analyser at 14MHz and adjusting the potentiometer for minimum IF feedthrough. The setting is quite sharp and ensures good mixer balance. An RF test signal of 11dBm (0.8V RMS) was used for each test signal for the two-tone IMD tests. The gate bias level chosen enabled an input level of +12dBm to be reached before the IMD increased sharply.

The performance of the H-mode test mixer was as follows. With an input RF test level of +11dBm (spaced at 2kHz or 20kHz) the conversion loss was 8dB; RF to IF isolation -68dB; LO to IF isolation -66dB. Input intercept points: 1.8 to 18MHz +53dBm; 21 to 28MHz +47dB, or better; 50MHz +41dB. These results were achieved with a gate-to-source DC bias of +1.95V and -8V substrate bias, a square-wave local oscillator amplitude of 9V and IF at 9MHz.

It seems likely that a good performance could be achieved with an H-mode mixer transformer-driven from a sine-wave source, provided the injection is via capacitors so that bias pots could still be used. There seems no reason why an H-mode mixer should not be used in an up-conversion arrangement rather than for a 9MHz IF.

RECEIVER PROJECTS

The Yearling

This easy-to-build receiver design (**Fig 6.105**) by Paul Lovell, G3YMP, was first published in *D-i-Y Radio* and later in *Radio Communication*. The original design was for the 20m band but it has been extended to cover 80m as well. The receiver is powered from either a 9V PP3 battery or mains adapter and can be built either on the printed circuit board (PCB) supplied with the kit or (with the help of an experienced constructor) on a prototype board. The building instructions here are based on the PCB and kit. Either headphones or a loudspeaker can be used.

Circuit design

Direct conversion was the first option considered. This type of receiver has the merit of simplicity but, unless carefully designed and constructed, can suffer from problems such as strong breakthrough from broadcast stations on nearby frequencies. Also, a stable VFO at 14MHz, while certainly not impossible, is not something to be undertaken lightly by a beginner. The superhet was then considered. This overcame some of the problems associated

Fig 6.105: The Yearling receiver

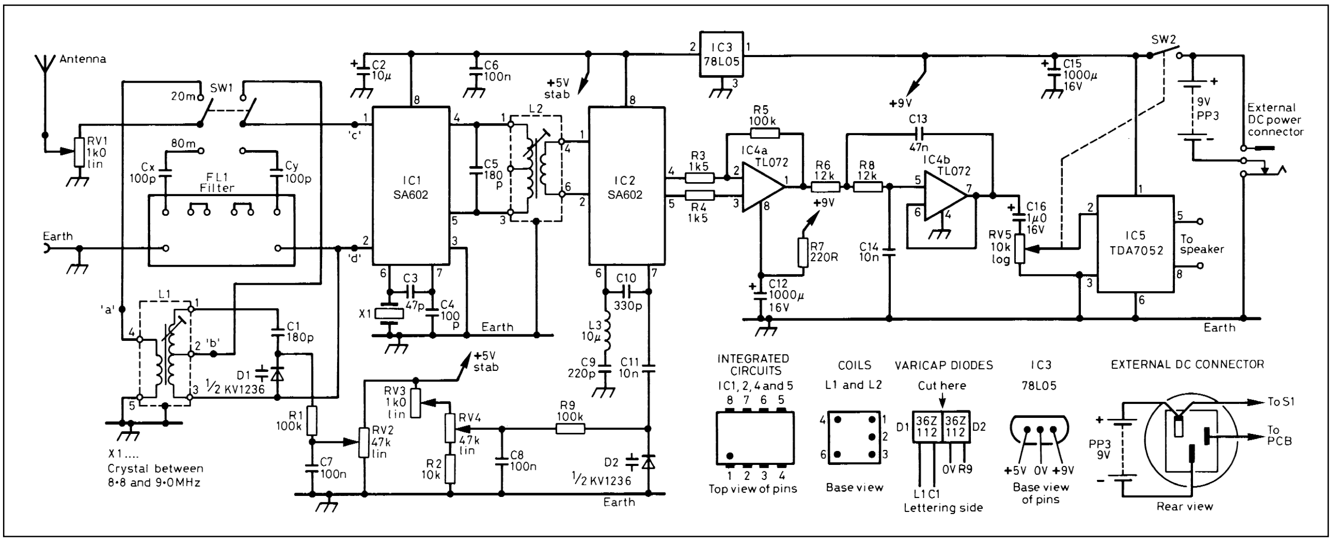


Fig 6.106: Circuit diagram of the easy-to-build Yearling receiver (see Table 6.5 for the Components List)

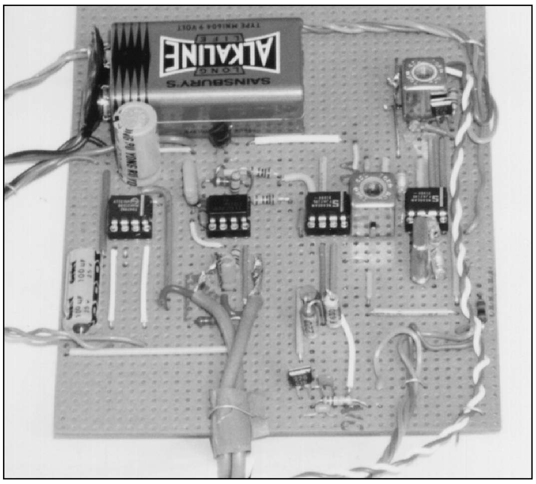


Fig 6.107: The Yearling may also be built on a prototype board

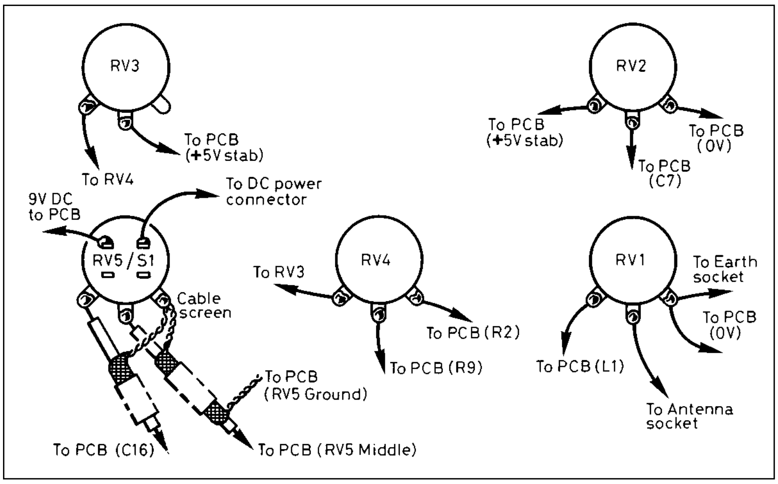


Fig 6.108: Rear view of the variable resistors. Check the connections carefully to make sure that the wires fit the correct holes on the board

with direct conversion but created others such as the need for a rather expensive IF filter. So the result was a happy compromise, which in effect is a direct-conversion receiver preceded by a frequency converter. This means that the VFO runs at about 5MHz instead of 14MHz, so stability is much better.

The circuit is given in **Fig 6.106**. Incoming signals at 14MHz or 3.5MHz are selected by the tuned circuit L1/C1/D1 or filter FL1. Note that the ANT TUNE control is not needed on 80m as FL1 provides the necessary filtering. Tuning on 14MHz is carried out by RV2 which adjusts the voltage on varicap D1. This is one half of diode type KV1236 - note carefully the polarity of this component.

A Philips SA602 (IC1) converts the signal to the range 5.0 to 5.5MHz (approx) by means of its internal oscillator. This has a crystal (X1) working at about 8.9MHz. In fact any crystal between 8.8 and 9.0MHz will be satisfactory but a frequency of 8.95MHz will give greatest accuracy on the dial. It will be noted that D1 is in fact forward biased over part of its voltage range. However, the circuit as it stands performs quite adequately.

The signal output from the mixer in IC1 then passes to the IF filter formed by C5 and L2. The tuned circuit is damped by the rather low output impedance of IC1, and this gives a nice compromise between selectivity and insertion loss. The balanced output of the tuned circuit is applied to IC2, another SA602 mixer/oscillator which acts as a product detector.

The main VFO uses the oscillator section of IC2, which covers a range of approximately 5.0-5.5MHz. Assuming the use of a 9MHz crystal for X1, the 20m band will track within the range 5.00-5.35MHz and the 80m band from 5.2-5.5MHz. Note that the LF ends of the respective bands will be at opposite ends of the dial, since 20m makes use of the sum of the receiver's two oscillator frequencies, and 80m uses the difference between them.

Main tuning is carried out by RV4, and RV5 provides the band-spread control. Tuning is by means of the voltage on varicap D2 which, in association with C9, C10 and L3, determines the frequency of oscillation. The varicap is a dual type - cut in two with a sharp knife. Voltage regulator IC3 in the supply lines to the early stages makes stability surprising good. The audio output from IC2 is amplified by IC4a and filtered by low-pass filter IC4b, before being further amplified to speaker level by IC5.

Construction

Many Yearlings have been built from kits without problems, while a number of constructors have successfully used a prototype board instead of the PCB. The prototype was built using just such a method as illustrated in **Fig 107**. No special precautions are needed but, as with any radio, neat wiring makes the tracing of faults a much easier process. **Fig 6.108** shows

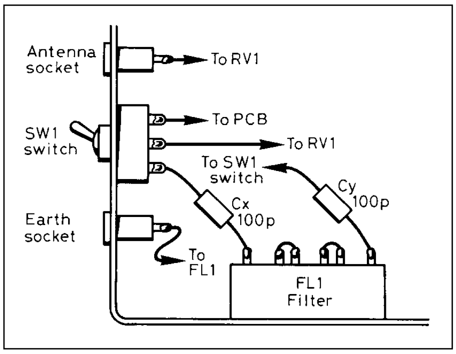


Fig 6.109: Internal view of the yearling case. The 80m filter is attached to the base with glue

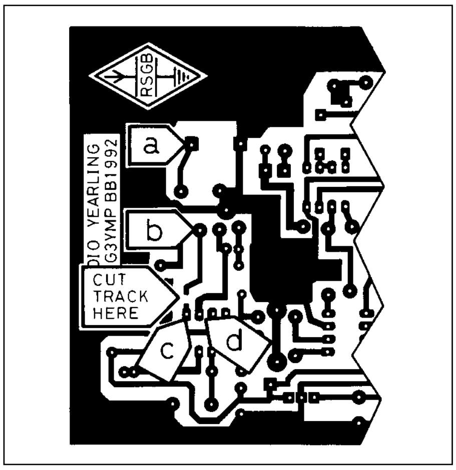


Fig 6.110: The underside of the PCB. Wires are connected from the switch and filter as shown

the connections to the gain and tuning controls. Screened cable should be used for the leads to the volume control but stranded bell wire should be satisfactory elsewhere. Incidentally, there was no problem with using IC sockets for all the 8-pin devices. The coils are colour coded, with L1 having a pink core and L2 a yellow one.

It is rather easy to wire the varicaps incorrectly but, if using the PCB, the lettering on D1 should be next to coil L1 and the lettering on D2 should be facing resistor R7. Fig 6.109 shows the band-change switch and the 80m filter which is glued to the side of the case. Holes for the five controls are 10.5mm diameter, and the speaker and power connectors have 6.3mm and 11mm holes respectively. The antenna and earth sockets need 8mm holes.

Setting it up

Connect a 9V battery and a reasonable antenna, and, on switching on, some stations - or at least some whistles - should be heard. It is suggested that a start is made with the 20m band, and the adjustments made before fitting the controls and sockets to the case.

A signal generator is useful, of course, but not essential to get the receiver working. The following steps should be followed, where the Yearling should burst into life.

1. Set the core of L2 to mid-position.
2. Set RV1, RV2 and RV4 to mid-position and rotate the core of L1 until you hear a peak of noise. Now adjust L2 for maximum noise.
3. Tune carefully with the main tuning control RV4 until you hear amateur signals. Adjustment of the bandspread may be needed to clarify the speech.
4. Switch off the receiver and fit the controls and sockets to the case.
5. Finally, adjust the tuning knob so that the pointer roughly agrees with the dial. Due to the spread of varicap capacitance values, you may find the tuning a little cramped. This is easily fixed by adding a resistor (try 22kΩ to start with) in series with RV3 or adjusting the value of R2.
6. Check the 80m band - this should work without further adjustments to the coils. Fig 6.110 shows the additional connections for 80m as the Yearling was originally designed for 20m only.
7. Finally, fix the PCB inside the case (double-sided sticky tape works well).

Capacitors (*All rated at 16V or more*

C1, 5	180p polystyrene, 5% or better tolerance
C2	10µ electrolytic
C3	47p polystyrene, 5% or better tolerance
C4, Cx, Cy	100p polystyrene, 5% or better tolerance
C6-8	100n ceramic
C9	220p polypropylene, 2% or better tolerance
C10	330p polypropylene, 2% or better tolerance
C11, 14	10n ceramic
C12, 15	1000µ electrolytic
C13	47n, 5% polyester
C16	1µ electrolytic

Resistors (*All 0.25W 5%)*

R1, 5, 9	100k
R2	10k
R3, 4	1k5
R6, 8	12k
R7	220R
RV1, 3	1k linear
RV2, 4	47k linear
RV5	10k log with switch (SW2)

Inductors

L1	Toko KANK3335R
L2	Toko KANK3334R
L3	10µ, 5% tolerance (eg Toko 283AS-100)

Semiconductors

IC1, 2	Philips SA602
IC3	78L05 5V 100mA regulator
IC4	TL072 dual op-amp
IC5	Philips TDA7052 audio amp

Additional items

Varicap diode, Toko KV1236 (cut into two sections)
Crystal, between 8.8 and 9.0MHz. An 8.86MHz type is available from JAB, Maplin, etc
Wavechange switch, DPDT changeover type
8-pin sockets for IC1, 2, 4 and 5
4mm antenna (red) and earth (black) sockets
3.5mm chassis-mounting speaker socket
DC power socket for external power supply (if required)
4 knobs, approx 25mm diameter with pointer
Tuning knob with pointer, eg 37mm PK3 type
Printed circuit board or prototype board
Plastic case, approx 17 x 11 x 6cm, eg Tandy 270-224.
Speaker between 8 and 32 ohm impedance (or headphones)

Kits of components & PCB are available from JAB Electronic Components, PO Box 5774, Birmingham B44 8PJ. E-Mail jabdog@blueyonder.co.uk

Table 6.5: Components list for the Yearling receiver

Conversion Loss:	1dB
Minimum Discernable Signal (MDS):	-136dBm
Two Tone Dynamic Range (2TDR):	111dB
Third Order Intercept (IP3):	+30dBm
Bandpass RF filter characteristic:	Q ~3000

Table 6.6: N7VE's performance figures for his commutative mixer

Experimental Direct Conversion Polyphase Receiver

This experimental design by Hans Summers, G0UPL, combines new and old techniques to produce a simple but high performance direct conversion receiver. As discussed above, the most obvious disadvantage of direct conversion is that both sidebands are detected. This can be solved via the use of audio phasing networks such that the unwanted sideband is mathematically cancelled by summation of correctly phased signals. Conventionally this required the use of two mixers, fed by phase shifted RF inputs.

A recent commutative mixer design by Dan Tayloe, N7VE, produces four audio outputs at 90-degree phase angles using a very simple circuit [20, 21]. Despite its simplicity the detector boasts impressive performance as shown in Table 6.6.

The circuit is not really a mixer producing both sum and difference frequencies: it might more accurately be described as a

BAND MHz	TYPE T1 - T2	Inductance T1 - T2	[pF] C1 - C3	[pf] C2
1.8 - 2.0	3333	45µH	150	12
3.5 - 3.8	3333	45µH	39	3.3
7.0 - 7.1	3334	5.5µH	100	8.2
10.1 - 10.15	3334	5.5µH	47	6.8
14.0 - 14.35	3334	5.5µH	22	3.3
18.07 - 18.17	3335	1.2µH	68	6.8
21.0 - 21.17	3335	1.2µH	47	4.7
24.89 - 24.99	3335	1.2µH	33	3.3
28.0 - 29.7	3335	1.2µH	22	3.3

Table 6.7: Details of using the receiver on all HF amateur bands

switching integrator. It possesses a very useful bandpass filter characteristic, tracking the local oscillator frequency with a Q of typically 3000.

The quadruple phased audio output of the Tayloe Detector is ideally suited to drive a circuit of much older heritage: the passive polyphase network. This consists of a resistor-capacitor network with resonant frequencies (poles) tuned such that they occur at evenly spaced intervals across the audio band of interest. The effect is a quite precise 90-degree phase shift throughout the audio band. In the current experimental receiver, the polyphase network connection is such that the unwanted sideband is mathematically completely cancelled inside the network, resulting in a

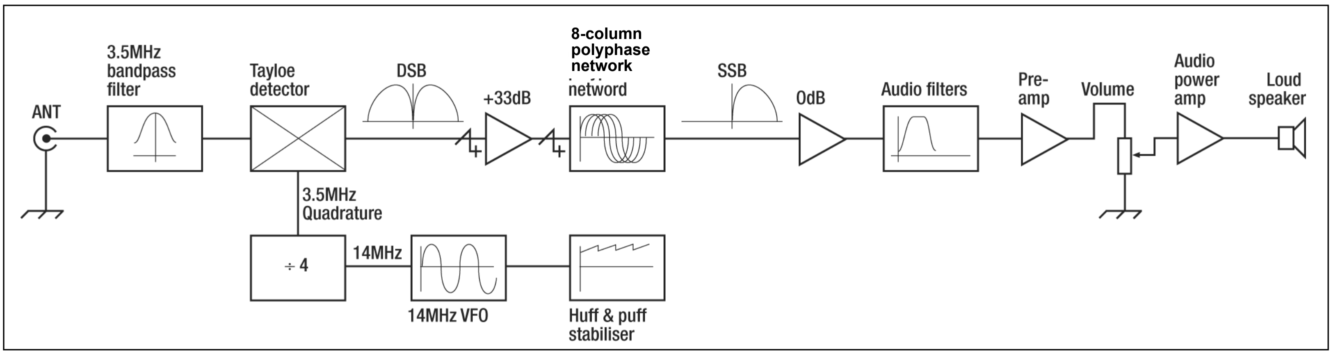


Fig 6.111: Block diagram of experimental Direct Conversion Polyphase receiver

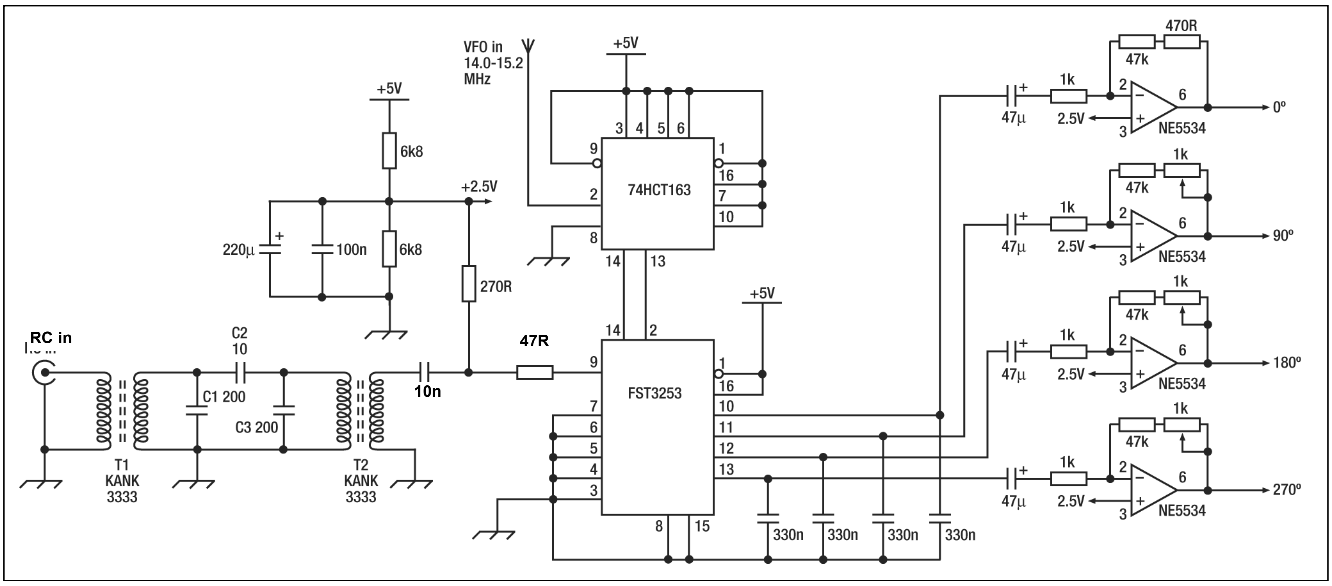


Fig 6.112: Front-end: RF filter and Tayloe Detector. Op-amps pin 4 is grounded and Pin 7 is +12V

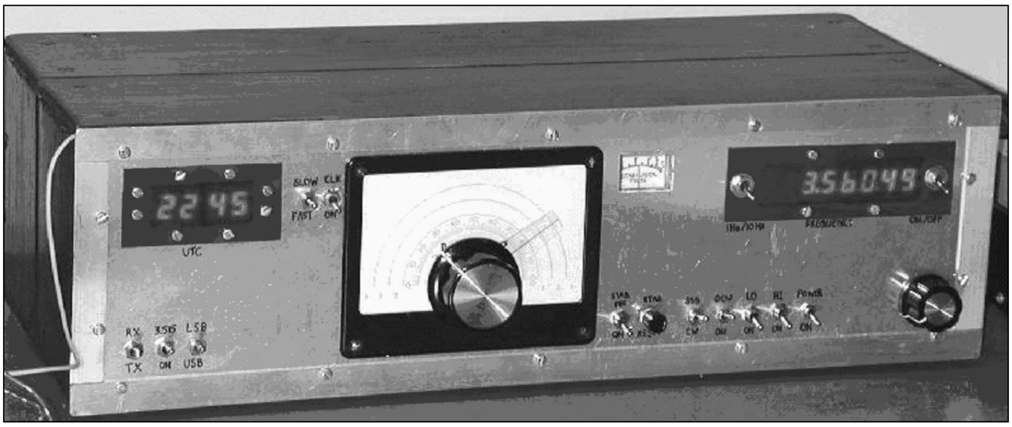


Fig 6.113: Prototype of G0UPL's experimental direct conversion polyphase receiver

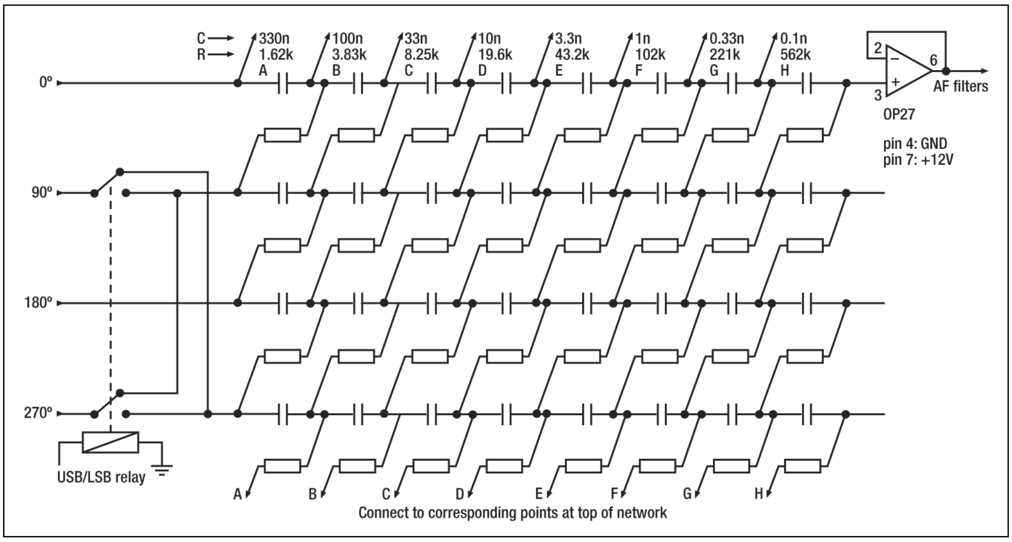


Fig 6.114: Polyphase network. Resistors and capacitors should be high tolerance types

single sideband audio output that is then amplified and filtered in a conventional way. **Fig 6.111** shows the block diagram of the experimental receiver design.

Input filter and detector

The high performance characteristics of the Tayloe Detector make a preceding RF amplifier unnecessary. In this receiver (**Fig 6.112**), the RF signal is filtered by a simple bandpass filter consisting of two TOKO KANK3333 canned transformers.

Circuit values are shown for 80m, but the circuit can readily be adapted to any HF amateur band, see **Table 6.7** from [22]. The input to the Tayloe Detector is biased to mid-rail (2.5V) in order to obtain maximum dynamic range. The FST3253 IC is a dual 1-4 way analogue multiplexer designed for memory bus switching applications and possessing high bandwidth and low ON-resistance. A local oscillator signal at four times the reception frequency is required to accurately generate the necessary switching signals: a synchronous binary counter type 74HC163 accomplishes this easily.

The four audio outputs are buffered by low noise NE5534 op-amps configured for a gain of 33dB. The gain of three of the op-amps is made adjustable by mult-turn 1kΩ preset potentiometers to allow the amplitude of each of the four paths to be matched precisely.

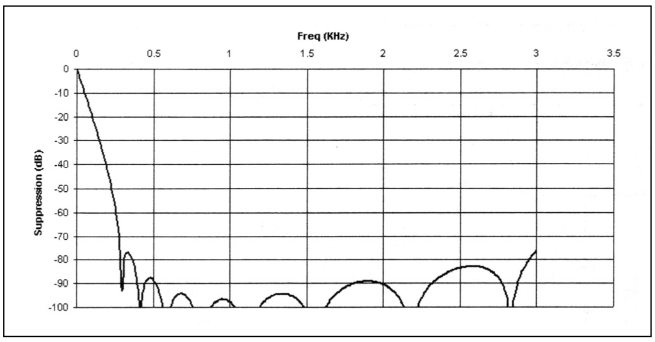


Fig 6.115: Theoretical opposite sideband suppression (dB) versus frequency (kHz)

Polyphase network

Polyphase networks are usually designed using either a constant capacitance value throughout the network, or constant resistance. Since capacitors are available in less-closely specified tolerances compared to resistors, a constant value capacitance made it possible to choose sets of four closely matching capacitors for each column in the network. However such a network can result in considerable losses, which are not constant across the audio band of interest. These losses must be compensated by higher amplification elsewhere in the signal chain, which generally degrades noise performance and dynamic range of the receiver.

This component values for this design were calculated by reference to an excellent article by Tetsuo Yoshida JA1KO [23]. Tetsuo's unique design process increases the resistance value from column to column in such a way as to produce a lossless passive polyphase network. This counterintuitive result has been verified by measurement.

An 8-column polyphase network was designed (**Fig 6.114**). The theoretical opposite sideband suppression of this network (**Fig 6.115**) assumes precise component values. In practice, this is impossible to achieve and the real world performance will be degraded somewhat compared to the ideal curve. This degradation can be mitigated as far as possible by the use of high accuracy (0.1%) resistors, together with 'padding' capacitors by connecting smaller capacitors in parallel until measurement indicates four matched capacitors for each column.

Note that selection of Upper or Lower sideband is simply a matter of swapping the 90- and 270-degree inputs to the network. This could be accomplished by a DPDT relay or switch; or alternatively for single-band use the circuit could be hard-wired. A single unity-gain high impedance low-noise op-amp (OP27) follows the polyphase network.

Oscillator

The Tayloe switching detector requires a VFO at four times the receive frequency. One way to obtain a stable VFO is by use of a Huff & Puff stabiliser. This simple circuit was developed initially by Klaas Spargaren PA0KSB in the 1970s and many subsequent modifications have appeared in *RadCom's* 'Technical Topics' column and elsewhere (see also the chapter on oscillators).

The stabiliser compares the VFO to a stable crystal reference frequency and locks the VFO to this reference in small frequency steps. It might be described as a 'frequency locked loop' rather than a 'phase locked loop'. The locking process is a slow loop, and lacks the complication of phase locked loops and frequency synthesisers, whilst easily achieving a low phase noise output.

Two simple, minimalist designs (**Figs 6.116 and 6.117**) developed recently by G0UPL make it even simpler to build a Huff & Puff stabilised VFO. These designs combine the VFO and stabiliser, resulting in a stable output frequency at TTL-logic levels, perfectly suited for driving the Tayloe Detector.

Fig 6.116 shows an implementation of PA0KSB's original stabiliser. This design uses one half of a 74HC74 D-type flip flop uniquely forced to behave as a simple inverter gate, and pressed into service as an oscillator. The crystal reference frequency is generated from a cheap 32.768kHz watch crystal. The frequency step size of the resulting VFO is determined by the division ratio of the 32.768kHz reference frequency, eg 32Hz. Remember that in the Tayloe Detector (Fig 6.107) the VFO is divided by four, which will also divide the tuning step by the same factor.

A later development was the 'fast' stabiliser by Peter Lawton, G7IXH, which was described in an article in *QEX* magazine [24]. He used a shift register as a n-stage delay line and compared the input and output of the delay line using an exclusive OR-gate (XOR). The effect is a statistical averaging of the output control signal. The 'fast' method makes it possible to stabilise a worse VFO compared to the standard method, or stabilise a comparable VFO with much less frequency ripple. The frequency step-size is given by:

$$\text{Step} = 10^6 \times \text{VFO}^2 \,/\, (z \times M \times \text{xtal})$$

where VFO is the VFO frequency in MHz, z is the number of stages of delay, xtal is the crystal reference frequency in MHz, and M = 2^n where n is the number of divide-by-2 stages in the VFO divider.

The minimalist design in Fig 6.113 uses only three ICs to implement G7IXH's 'fast' Huff & Puff method. One XOR gate is used as the VFO. The shift 74HC164 register effectively provides a 7-stage delay line.

To increase this further, a 4517 CMOS IC (128-stage shift register) could be cascaded in series with the 74HC164 to provide a 135-stage delay line (**Fig 6.118**). Note that the 4517 (part numbers HEF4517, CD4517 etc) is a member of the original CMOS 4000-series and was not produced in later, higher speed families such as the 74HC-series.

Therefore it must be connected AFTER the 74HC164 so that the '164 is responsible for detecting the fast edges of the 60MHz reference oscillator.

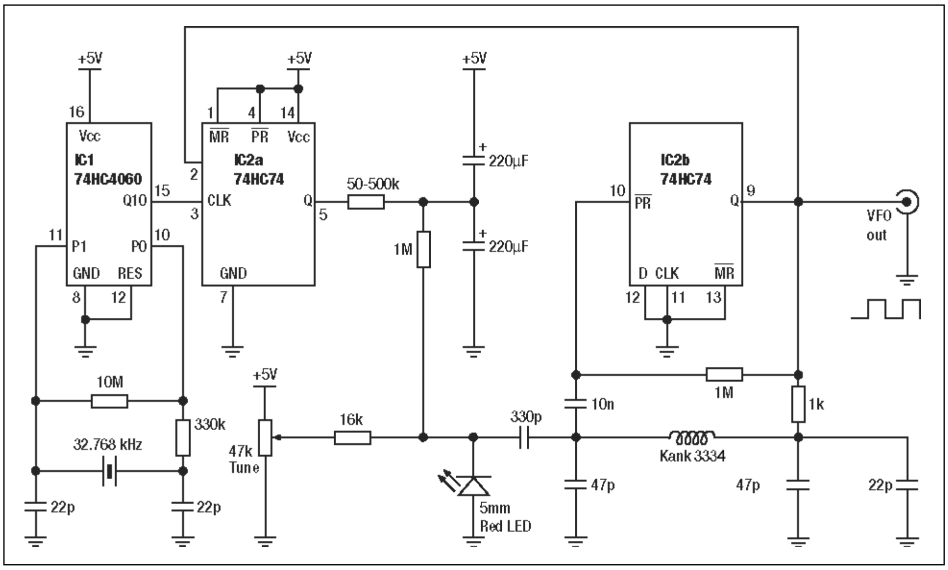


Fig 6.116: Two-chip simple Huff & Puff stabilised VFO

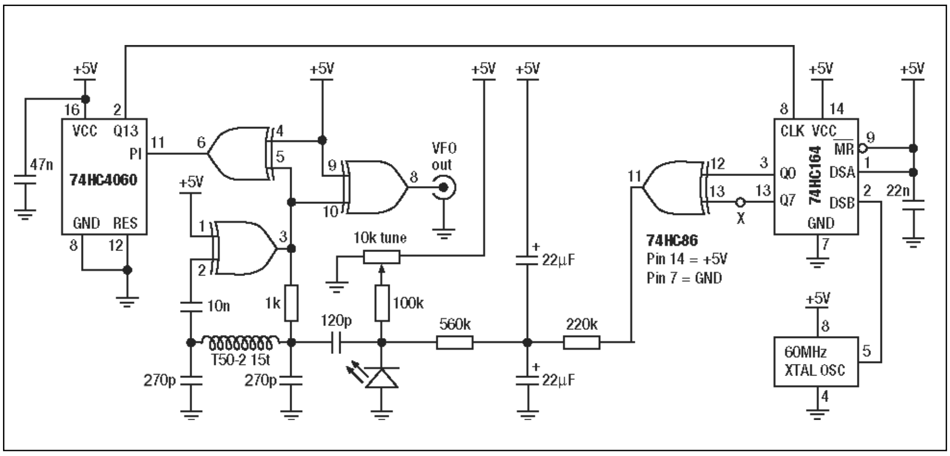


Fig 6.117: Three-chip 'fast' Huff & Puff stabilised VFO, offering improved performance

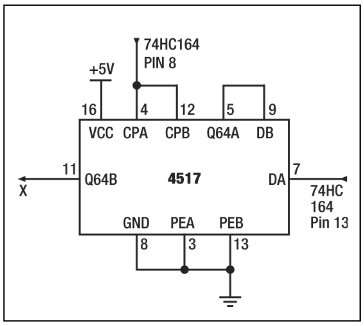


Fig 6.118: For higher frequencies, a 4517 CMOS IC may be added to improve the Huff & Puff stabiliser even further

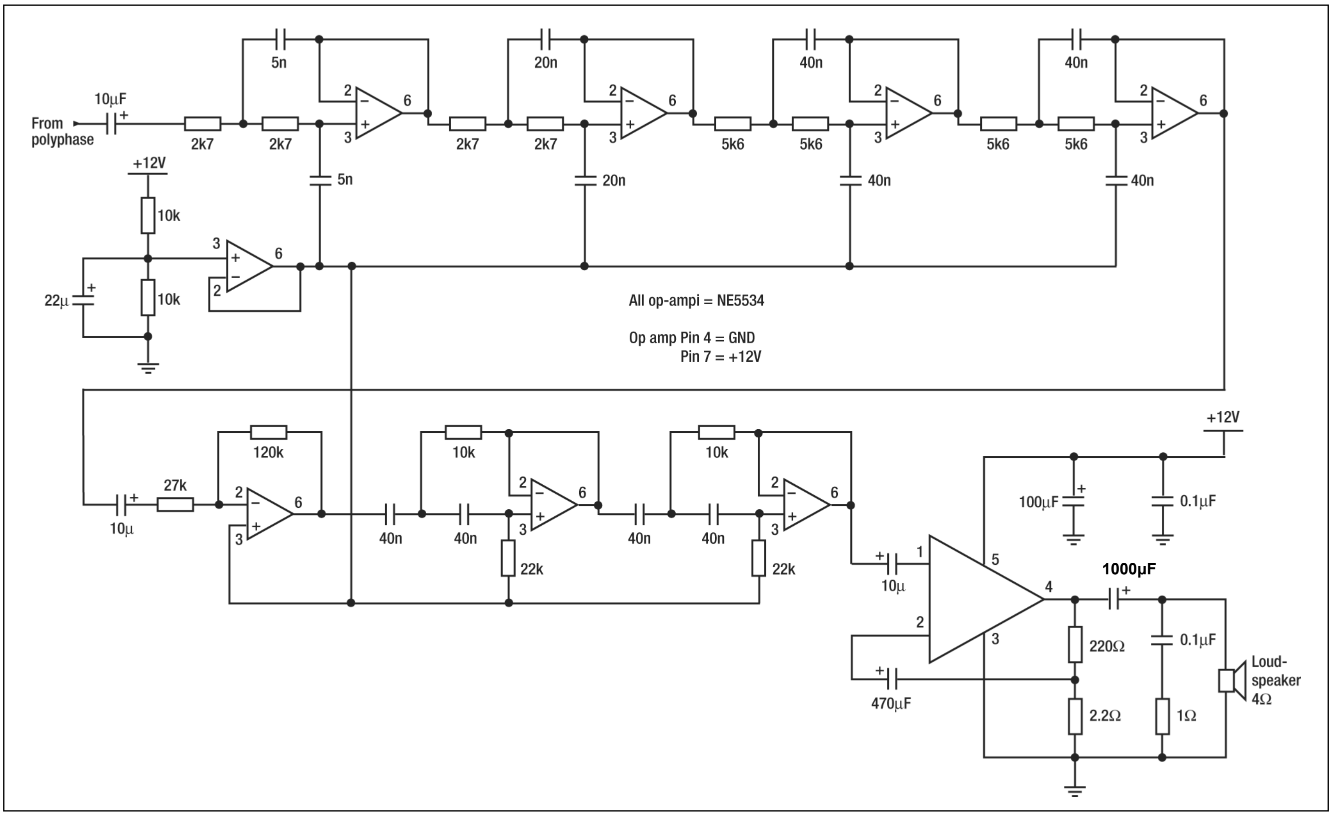


Fig 6.119: Audio stages: Pre-amp, 8-pole low pass filter, 4-pole high pass filter, and TDA2002 power amplifier

This 'fast' design is recommended for higher frequency VFOs such as might be used to build this experimental receiver for higher HF bands.

Audio stages

The remainder of the experimental receiver is relatively conventional and non-critical. Low-noise NE5534 op-amps are used to construct high-pass, low-pass filters to restrict the SSB bandwidth to 300Hz - 2.8kHz. A switchable narrow filter at 800Hz could be added for CW reception. A standard TDA2002 audio power amplifier, produces sufficient output power to drive a 4Ω loudspeaker for comfortable 'arm-chair' copy. An example audio section is shown in **Fig 6.119**.

Conclusions and further development

The receiver described has been found very satisfactory in use. Other designers of similar receivers have implemented parts of the circuit slightly differently. Many of these modifications would make the circuit more complex, but the following points might suggest avenues for further experiment:

1. The switching order of the divide-by-4 circuit in the Tayloe detector can be altered to a 'gray code' sequence such that only one of the 2-bit outputs changes state at each clock pulse. This is said to produce lower switching noise, though on 80m the atmospheric noise probably swamps any such effects anyway.
2. A clock-squarer circuit can be employed to generate a precise 50% duty cycle from a VFO at two times the reception frequency; this obviously imposes less stringent requirements on the VFO, which is harder to construct for higher frequencies.
3. The VFO could be replaced by one generated by Direct Digital Synthesis or other precise oscillator methods (see the Building Blocks: Oscillators chapter).
4. It is possible to use the other half of the FST3253 switch in parallel to halve the ON-resistance; alternatively a double-balanced Tayloe detector may be constructed using the second switch and a phase-splitting transformer at the input. This would make the detector more immune to noisy VFO signals such as might be produced by digital methods, eg DDS.
5. Many designers combine 0, 180 and 90, 270 degree outputs of the Tayloe detector prior to feeding the polyphase network. This can reduce certain common-mode noise sources.
6. Instead of using just one output from the polyphase network, all four can be combined thereby averaging errors and improving the signal-to-noise ratio.
7. For transmit operation, it is possible to connect the Tayloe Detector in 'reverse' as a high-performance SSB modulator to make a simple, high performance direct conversion SSB transceiver.

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