Direct Conversion Transceiver Design for Compact Low-Cost Portable Mobile Radio Terminals

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Abstract

The full benefit of using linear modulation methods for the transmission and reception of data in the mobile environment will only be realised when compact and power efficient linear transceiver equipment becomes a commercial reality. Moving from superheterodyne to direct conversion transceiver architectures is seen as a major step towards achieving this goal. Several characteristic problems with direct conversion technology have however hindered progress in this area.

This paper describes methods for overcoming many of the difficulties associated with direct conversion architectures and presents results from a working prototype developed by the Centre for Communications Research at Bristol University.

Introduction

The availability of high quality mobile telephone systems has proven to be an invaluable resource in many facets of business and social life. Yet, despite the economic benefits of a mobile phone, widespread acceptance has been hampered from the user perspective by the short battery life, terminal size and cost, and from the operator perspective by the shortage of spectrum, hindering service expansion. The spectral efficiency of radio communication systems is central to the future of mobile communications industry. It is predicted that in the UK alone, the present cellular telephone population of 1.5 million users will rise to 10 million by the year 2000 [1].

Relatively inefficient use of spectrum by the present FM system has manifested itself as crowding on the available channels. This is particularly prevalent in cities and on sections of motorways at times of peak demand. Overcrowding in the present UK TACS system results in long delays for connection and (occasionally) call termination. Battery life, whilst of less importance in vehicle-based equipment, has proved a decisive factor in the serviceability of hand held units. Present hand-held designs are a compromise between transmission range and battery life, and their performance falls short of similar vehicle-based transceivers. Also, even the smallest hand held units require a larger than average pocket.

Future public land mobile telecommunication systems (FPLMTS) will be required in both vehicle-based and hand-held form. The challenge of meeting the needs of power and spectrum efficiency in a compact package is a hard one. However, if any FPLMTS is to be truly universal and widely available, the eventual goal of a pocket-sized "personal communicator" must be realised. This has been recognised by several working groups [2,3], and a target of 200g weight and $200cm^3$ volume has been suggested [4]. There is also some consensus that a FPLMTS should be based on low bit-rate (16 kbits/s or less) voice technology. The transceiver systems employed in any FPLMTS must be power efficient and suitable for use with data modulation. In addition, the modulation scheme adopted must be as efficient in its use of spectrum as possible.

Previous work at the Centre for Communications Research at Bristol [5,6,7] has shown the arguments for a linear modulation (LM) scheme being one of the most spectrally efficient. Other work [8,9] has supported these arguments. Yet existing LM equipment of sufficient quality is bulkier at present than comparable FM equipment, and more power hungry. To utilise the obvious advantages of LM, a departure has to be made from conventional transceiver designs. A *direct conversion* transceiver system eliminates many of the space critical components found in superheterodyne designs, and allows the engineer much scope for integration. There are, however, severe problems with this approach [10,11]. This paper shows how one type of direct conversion transceiver has been developed at Bristol to the point where a prototype is under evaluation. Test results are provided to show the performance of this transceiver to be equivalent to (and in some cases better than) existing systems.

The Transmitter

The requirements for a transmitter in a LM system are that it should have adequate output power, low out-of-band emissions and (if possible) be power efficient. The last two attributes are incompatible in conventional transmitter power amplifier designs. Linear amplifiers generally have large quiescent currents, and efficient Class C amplifiers are highly non-linear.

One method of meeting all the above requirements is the

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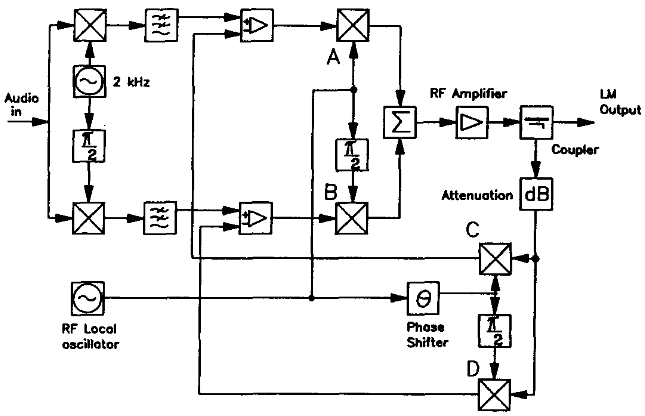

Figure 1: Cartesian Loop Transmitter with Phase Shifter

Cartesian Loop transmitter [12,13]. Figure 1 shows the entire transmitter in a Weaver method [14] configuration. This method was adopted as it allows image suppression which is superior to other techniques. It is far from being a new idea, but its development has been limited in the past because of three problems; loop stability, generation of accurate I and Q signals and maintenance of quadrature on the local oscillators.

These problems have been overcome by several developments. Firstly, loop stability is governed primarily by the phase shift needed between mixer pair AB and mixer pair CD. This phase shift corresponds to the delay through the power amplifier. Maintenance of the correct phase shift is therefore essential for spurious-free operation. A voltage-variable r.f. phase shifter θ has been incorporated into the design. This enables the relative phase of the two sets of local oscillators to be varied over a wide range, and also over a wide range of frequencies.

A related network has been developed to provide accurate quadrature for each set of local oscillators. Fine adjustment is possible by means of a voltage level, and quadrature can be maintained over a large bandwidth with minimal amplitude fluctuation.

Finally, a digital signal processing (DSP) system enables the production of very accurate I and Q baseband signals. This also allows control and calibration of the transmitter, during an initial calibration period and subsequently during operation.

The results of these developments are dramatic (Figure 2). An existing LM amplifier, biased in Class A or AB mode is palpably worse than the linearised Class C amplifier. Output powers of 20W PEP have been obtained at VHF with excellent suppression of spurii up to the bandwidth of the differential amplifiers. This may be termed the ***linearising bandwidth.*** This is different from the transmitter's ***bandwidth of operation***, which is dependent upon the quadrature local oscillator networks, the phase shift network and the amplifiers. This can be very wide (several tens of MHz). There are, however, limits on the Cartesian Loop Transmitter's linearising bandwidth, governed primarily by the phase delay within the feedback control loop. With too wide a loop bandwidth or too high a loop gain, instability will occur. Linearised bandwidths of several hundred kHz have so far been achieved with this technique.

The above results have been reproduced at UHF (900 MHz). The nature of components at these frequencies means

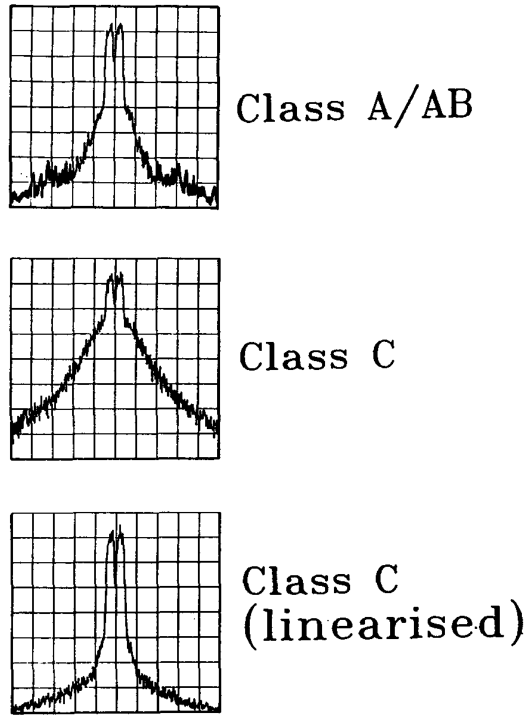

Figure 2: Outputs from different VHF amplifier types with a 4 kHz notch-filtered noise modulating signal

that the resulting transmitter may be even smaller in size than its VHF counterpart.

The gains in power efficiency are equally remarkable. Whereas a conventional LM amplifier (for 20W output) may have a quiescent current of about 100mA at 12V, the corresponding current in a linearised Class C amplifier is zero. LM is also generally more power efficient than FM, as Figure 3 shows. Here an FM system with a Class C power amplifier is compared with an LM system with a linearised Class C amplifier. Where the modulating signal has a high peak-to-mean ratio (such as speech) the overall gain in power efficiency is impressive.

A final advantage is that the Cartesian Loop Weaver method transmitter uses components which may all be reduced in size to chip or hybrid level integration.

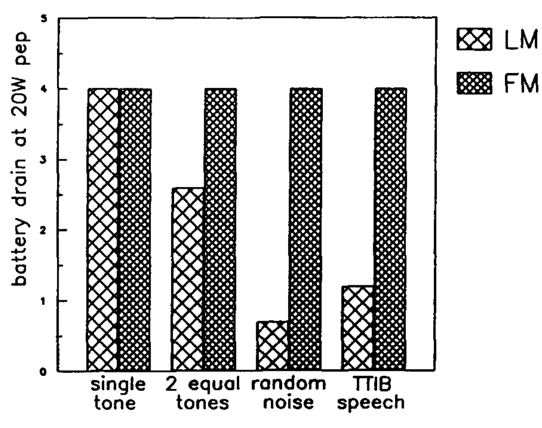

Figure 3: Comparison of battery drain (in amps) for an FM Class C amplifier and for an LM linearised Class C amplifier with different modulation types

The Receiver

A Weaver architecture has also been employed in the receiver section. There are severe difficulties inherent in this type of receiver [15], but they have largely been solved by careful design and advanced DSP techniques. These difficulties will be considered together with the requirements for sensitivity, dynamic range and carrier feedthrough.

Sensitivity

Determining the sensitivity requirement for a receiver is not straightforward. It is desirable to have system sensitivity as high as possible, but not to the point where other aspects of receiver performance are degraded, particularly strong signal handling. A guide to necessary sensitivity may be obtained by looking at two comparable radio sets; an Aerotron ACSB Pioneer 1000 system (Set 1), and a Securicor T530 FM Mobile (Set 2). Their relative sensitivities (for 12 dB SINAD) are given in Table 1. From these values, it is clear that a sensitivity of around −120 dBm is necessary [18].

The front-end of the direct conversion receiver is shown in Figure 4. This compares with the front-end of a superheterodyne (superhet) receiver shown in Figure 5. A noise figure summary for a typical superhet receiver is given in Figure 6, for a sensitivity of −120 dBm (noise figure of 7.2dB). The top portion of the figure shows noise temperature and gain data for the individual stages. The bottom portion shows how the total noise figure of the receiver is built up from the individual stage data. Equations for deriving noise figures in cascaded stages are well-documented elsewhere [19].The output of the IF strip is used as a reference — it is assumed that after this stage little noise will be added. A corresponding noise figure analysis for the direct conversion receiver is shown in Figure 7. Here the input to the A/D converter is used as a reference.

This analysis shows that the direct conversion receiver relies heavily upon the preamplifier for even this modest sensitivity requirement. The presence of the signal splitter and passive mixers at such an early point in the signal path is undesirable. However, their inclusion (instead of active mixers) was necessary for improved intermodulation performance. The analysis also demonstrates that the audio amplifiers are pushed to the limits of their noise capabilities.

The required sensitivity can just be achieved. Future work will include the evaluation of higher quality active mixers and a very low noise audio preamplifier to enhance sensitivity.

<table>
<tr><td colspan="2">Transceiver</td><td>Sensitivity (dBm)</td></tr>
<tr><td rowspan="2">Aerotron 1000</td><td>with preamplifier</td><td>-125</td></tr>
<tr><td>without preamplifier</td><td>-119</td></tr>
<tr><td colspan="2">Securicor T530</td><td>-120</td></tr>
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Table 1: Sensitivities of two commercial PMR transceivers

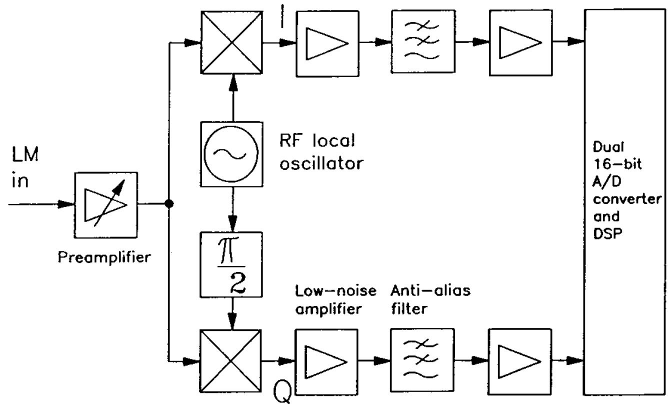

Figure 4: Digitally-implemented Weaver Direct Conversion receiver

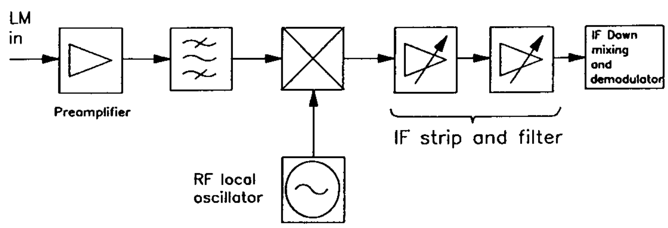

Figure 5: Single-conversion Superhet Receiver front-end components

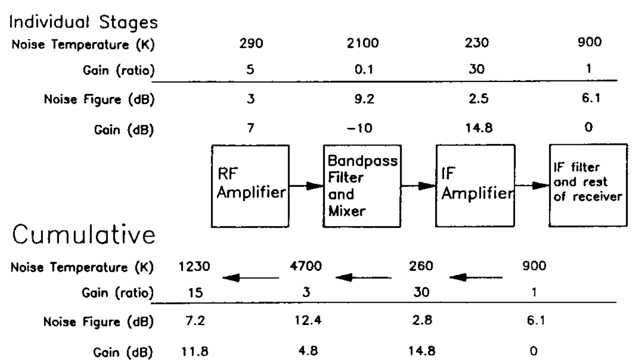

Figure 6: Superhet receiver noise figure analysis

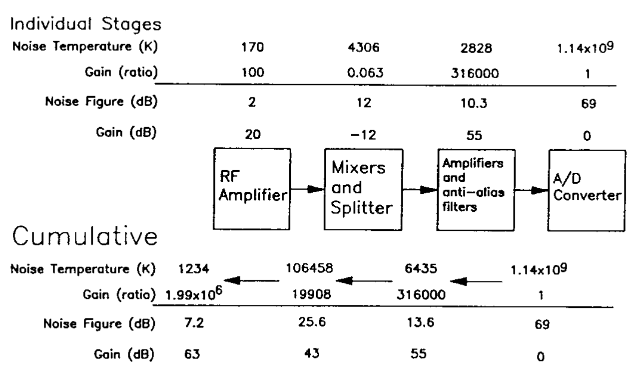

Figure 7: Direct Conversion receiver noise figure analysis

	Component			
	Preamplifier	Mixers	Amplifier/Filter	A/D Converter
Maximum Signal (dBm)	−20	−20	+23	+23
Minimum Signal (dBm)	−120	−130	-115	-60

Table 2: Dynamic ranges for Direct Conversion Receiver components

Dynamic Range

There are two aspects to the dynamic range of a receiver. Firstly, there is the range of signals in the wanted band that the receiver is required to tolerate. This may be as high as 120 dB. Secondly, there is the largest ratio of adjacent signal to wanted signal that can be processed before third-order distortion products swamp the wanted signal. This might reasonably be 70 to 80 dB [16] and is termed the spurious-free dynamic range (SFDR). Into this area also comes the question of adjacent channel rejection and selectivity.

SFDR will be considered first. It is affected by every component in the signal path. Referring to the direct conversion receiver front-end (Figure 4), each component can be assigned a dynamic range. That is, it can be given a range of signals which can be amplified without distortion (at the high end) or without disappearing into noise (at the low end). These are summarised in Table 2. Figure 8 shows how an incoming signal range from −120 dBm to −40 dBm can be translated to −67 dBm to 23 dBm for a 16-bit analogue to digital (A/D) converter. SFDR is directly applicable to A/D converters as well. The signal-to-noise (S/N) ratio of an n-bit A/D converter is given by [17]:

$$\left[\frac{S}{N}\right]_{dB} = 4.8 + 6n$$

For a 16-bit converter, the S/N ratio is approximately 96 dB. This does not take account of thermal noise. It also does not correspond to dynamic range. The smallest signal that may be accomodated 12 dB above the noise floor is $96 - 12 = 84$db below the full-scale signal. This assumes a perfectly ideal converter. Until recently, 16-bit converters at reasonable cost achieved, in practise, only 14 or 15-bit resolution. The advent of self- calibrating converters which provide true 16-bit resolution with very low differential non-linearity has made the construction of this receiver possible. Future oversampling converters which require minimal anti-alias filtering should enable the receiver to be simplified further. The point has now been reached where the linearity of the receiver is determined by other receiver components, in particular the audio amplifiers.

The direct conversion receiver developed uses digital signal processing (DSP) to perform several tasks usually undertaken by analogue circuitry. Adjacent channel filtering, automatic gain control (AGC), squelch control and demodulation are all carried out on one DSP chip. In particular, digital adjacent channel filtering is a significant advance as it allows linear phase low-pass filters to determine the channel characteristic. Existing LM (and FM) equipment suffers from the inclusion of a crystal IF filter. The gain and group delay characteristics of a typical 10.7 MHz IF filter, compared with similar characteristics for a digital filter, are shown in Figure 9. It is clear from these graphs that the digital filtering provides a much better behaved channel characteristic, particularly suited to data transmission.

In addition, digital channel filtering provides well-defined

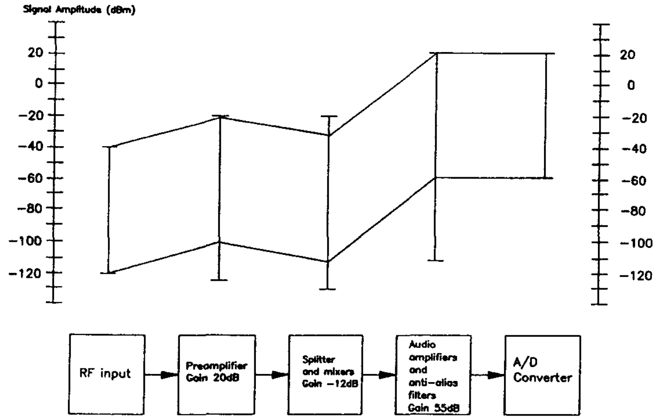

Figure 8: Translation of dynamic range from input to output in the Direct Conversion Receiver

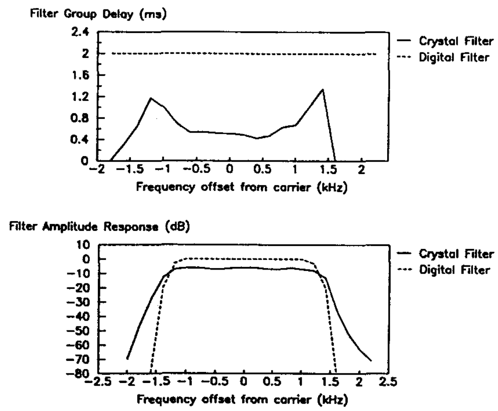

Figure 9: Comparison of Gain and Group Delay Characterisitcs for Digital and Crystal Channel Filters

adjacent channel performance. Figure 10 shows the selectivity[1] curves for the direct conversion receiver, set 1 and set 2. From these curves, it is clear that the direct conversion receiver has both a tighter and better defined selectivity performance than either of the other radios.

The direct conversion receiver has, as required, an SFDR of about 80 dB. This SFDR may be located between 0 and −120 dBm by varying the gain of the preamplifier. Under software control, the gain of the preamplifier may be adjusted from −20 to 20 dB, and thus the overall dynamic range of 120 dB is achieved.

[1]The input signal level is set to give 12dB SINAD. An interfering signal is then introduced at a variable frequency offset, and its level is adjusted to degrade the wanted signal to 6dB SINAD

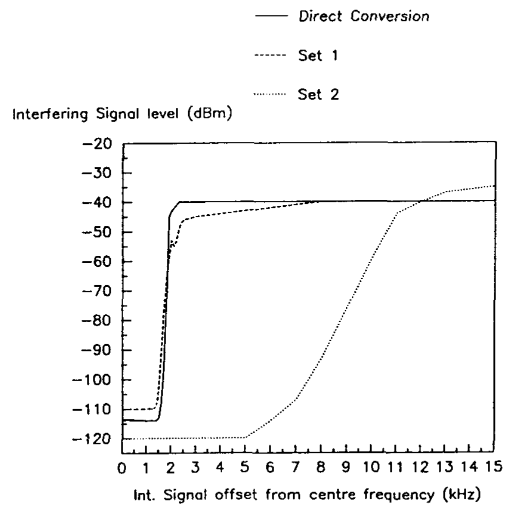

Figure 10: Adjacent Channel Rejection responses for the receivers under test

A related problem is that of gain and phase matching in the I and Q signal paths. Accurate quadrature of the local oscillators is maintained by a network similar to that outlined in the transmitter section. The amplitudes of the I and Q signals are compared in software and any required adjustments made on a continuous basis. These techniques have the effect that image suppression of 40dB or better is maintained at all times.

Carrier Leakage

Carrier leakage and feedthrough onto the I and Q signal paths results in a d.c. level being present. Also, the audio amplifiers and anti-alias filters may contribute some d.c. offsets. If fed through to the second mixing stage, this d.c. level appears as an undesirable tone in the centre of the band.

Three methods of addressing the carrier leakage/d.c. offset problem have been identified [15]. The most promising of these is d.c. correction — a software method whereby the incoming signal is averaged over a relatively long period, and the result subtracted from the signal. This method has been implemented, and works to a large extent. It is a substantial improvement over a.c. coupling for two reasons. If the I and Q paths are a.c. coupled at (say) a cut-off of 50 Hz, then a significant amount of information is lost in the notch created [10]. If the coupling is reduced to 5 Hz, the notch is narrower, but the group delay characteristic of the a.c. coupling filter adversely affects the channel characteristic. The DC correction technique differs from the above in that it is essentially a discrete system, correcting for d.c. error only at specific instants in time. The result is that the group delay response of the receiver is thus left largely unaffected. A notch is still created, but it can be made very narrow without introducing significant delay distortion. In a pilot-based system, such as TTIB, where the pilot is nominally in the centre of the frequency band (and thus appears at d.c. in a Weaver demodulator), the effects of the d.c. nulling notch can be overcome by introducing a small frequency offset (10 Hz or so) into the downconverted signal.

The d.c offsets on audio amplifiers and anti-alias filters are unfortunately not constant, and even very small variations have a significant effect if the incoming signal is also very small. Carrier leakage too is not constant, depending on environmental and circuit effects. Correction for d.c. offsets therefore has to occur at a rate sufficient to counteract the change. At present, correction is used in the direct conversion receiver at approximately 2–3 times per second. Careful front-end redesign and the use of low-drift amplifiers should allow this rate to be reduced at least ten times, with the result that an effective notch width of < 1 Hz is experienced. Other techniques, including local oscillator dither to reduce the d.c. content of the down converted input are under investigation. A combination of d.c. correction and dither may well prove to be the most practical solution.

Receiver Fading Performance

The test arrangement for assessing the direct conversion receiver's performance in a fading environment is shown in Figure 11. The test signal used was a sine wave, for simplicity. The SINAD results are shown in Table 3. Set 1 was employed, with its preamplifier connected. This demonstrates the superior sensitivity of Set 1, and the superior strong-signal handling of the direct conversion receiver. Preliminary tests with speech indicate similar conclusions. It is expected that future data trials will demonstrate the advantage of the linear phase channel characteristic.

Receiver		SINAD measurements		
	Signal Strength (dBm)	No fading	Moderate fading	Severe fading
Set 1	-27	18	19	9
	-47	23	21	12
	-77	23	21	12
	-107	18	16	10
Direct conversion receivers	-27	22	21	10
	-47	22	21	10
	-77	22	20	10
	-107	17	14	8

Table 3: Comparative SINAD measurements for different levels of fading

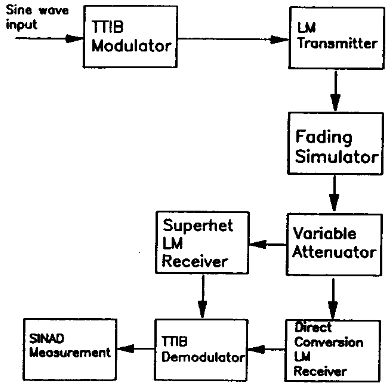

Figure 11: Test arrangement for fading comparison tests

Conclusions

The results presented in this paper show that a direct conversion transceiver system can be made to work at both VHF and UHF. The improved channel characteristics facilitated by digital channel filtering make this configuration particularly suited to data transmission. In addition, the integration potential of all the system components is the best available route to the concept of the 'personal communicator'.

Future work in this area will concentrate on techniques for automatic calibration and testing of the transmitter and receiver, the research and development of improved linear RF component technology, together with further evaluation of the prototype operation and reliability.

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