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Linearisation of frequency-hopped transmitters using Cartesian feedback

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Abstract This paper investigates the potential of the Cartesian feedback technique in linearising *frequency-hopped* (*FH*) transmitters for use in third generation mobile communication systems. Based on the results obtained from a transmitter hardware prototype, the practical problems resulting from a hopping carrier are discussed, and solutions to overcome such problems are presented.

1 Introduction

Both frequency hopping spread spectrum [1] and linear modulation techniques [2, 3] offer very attractive options in providing the high degree of spectral efficiency required for future mobile communication services. Clearly, the combination of the two techniques can result in an even higher spectral efficiency than that achievable using either FH or linear modulation alone.

A *linear transmitter* with a wide enough operating bandwidth to accommodate the range of hopping carrier frequencies is the very first requirement for an FH linear system. The effectiveness of the Cartesian feedback technique in the linearisation of radio frequency (RF) power amplifiers [4-7] suggests its suitability for such a system.

To investigate the properties of the ***FH Cartesian transmitter***, a hardware prototype operating in 215-225 MHz range was constructed at the Centre for Communications Research, University of Bristol. Using this model, issues such as transmitter transient response, both at baseband and RF, effects of variations in the amplifier non-linear characteristics and the loop parameters with frequency on the transmitter, and the required control signals for successful operation were closely examined. The outcome of this investigation is presented in this paper.

2 Frequency-hopped transmitter

A general block diagram of the FH transmitter is shown in Figure 1. As with the basic single-carrier version, the baseband input signal is applied to the two arms of the transmitter in an I-Q format. A fraction of the RF output signal is applied to the input of the down-converter. Suitably scaled, the down-converted signal is then subtracted from the pure input signal. The error signal thus generated is filtered by first-order low-pass filters to remove its high frequency components, and the result is up-converted, pre-amplified and finally applied to the RF amplifier.

If the total loop gain is sufficiently high and the feedback components do not add significant distortion, the result is a dramatic reduction in the distortion generated by the RF amplifier. The level of reduction depends on a set of parameters such as loop gain, bandwidth and delay, non-linear characteristics of the RF amplifier, and imperfections of the loop feedback components [7].

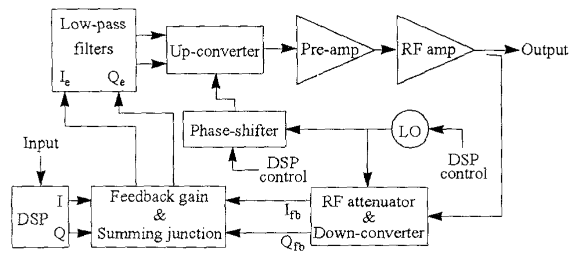

Figure 1: *The block diagram of the frequency-hopped Cartesian transmitter*

2.1 Parameters affected by hopping

The parameters which might change in the hopping process are the amplifier phase and gain (due to changes in frequency response), its non-linear characteristics, and the feedback gain/phase imbalance. The effects of these changes on the transmitter performance are discussed in this section.

Amplifier frequency response and non-linear characteristics

If the change in the amplifier gain with frequency is negligible, and if its phase response is linear within the transmitter hopping bandwidth, the linearising action of the transmitter at each carrier frequency will depend only on the amplifier AM-AM and AM-PM characteristics. If the amplifier non-linear characteristics remain the same at the different operating frequencies, the transmitter performance will stay consistent, and similar degrees of linearisation will be achieved throughout the hopping bandwidth.

Assuming a linear amplifier phase response, its baseband equivalent, representing the variations of the amplifier phase response with the frequency of the components of the modulating signal ω_b, is given by (1). In this equation, $\Phi(\omega_b)$ is the linear approximation to the phase response, $\Psi(\omega_c)$ the phase at the carrier frequency, and τ the absolute value of the gradient of the straight line approximating the phase response, estimating the amplifier group delay. Thus the baseband equivalent of the amplifier phase response for each operating frequency may be approximated simply by inserting the appropriate value of $\Psi(\omega_c)$ into (1).

$$\Phi(\omega_b) = \Psi(\omega_c) - \omega_b \tau \quad (1)$$

Since the linearising performance of the transmitter is determined by the level of its loop gain, and its stability by its gain-bandwidth-delay product [7], the value of loop time delay, determined mainly by the RF amplifier group delay, should remain relatively constant (i.e. requirement for a linear phase response) if consistent performance is to be achieved throughout the hopping process.

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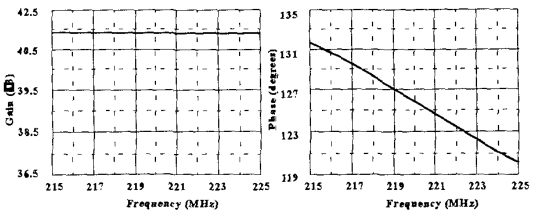


Figure 2: *Measured frequency response of the class AB amplifier used in the hardware prototype*

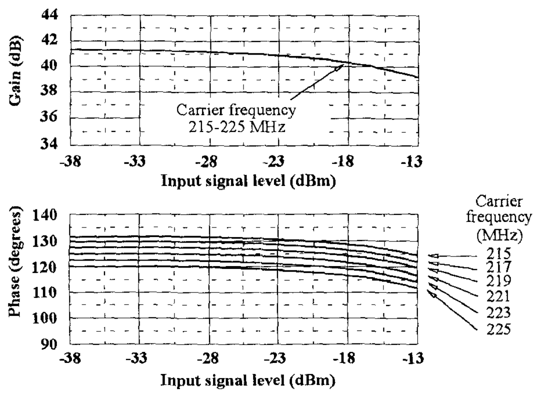


Figure 3: *Measured non-linear characteristics of the amplifier*

Transmitter stability will be seriously affected by large values of $\Psi(\omega_c)$. Hence it is necessary to compensate for this phase by appropriate phasing of the up- and down-conversion stages. As $\Psi(\omega_c)$ is dependent on the value of the carrier frequency, it is essential to synchronise the conversion stages dynamically, i.e. by generating the correct phase compensation for each hopping frequency. A voltage-controlled RF phase-shifter, which can be easily controlled by a digital signal processor (DSP), may be used to generate the correct phasing for the FH transmitter.

The spectral asymmetry caused by the amplifier AM-PM characteristic can also be reduced by further adjusting the phase-shifter generated phase. The value of this additional phase is determined by the amplifier non-linear characteristics. Thus, the voltages corresponding to the values of the phase shift required at each carrier frequency may be kept in a look-up table which is then used by the DSP to control the phase-shifter in a suitable manner.

Feedback gain/phase imbalance and DC offset [7]

The gain/phase imbalance and DC offset generated by imperfections in the feedback modules can degrade the transmitter performance severely. However, the effects of these imperfections may be significantly reduced or even removed by pre-distorting the transmitter input signal.

The main sources of feedback gain/phase imbalance and DC offset are the imperfections of the down-converter mixers. Naturally, the change in the carrier frequency during the hopping process affects the level of distortions generated by these imperfections, which in turn necessitates a different set of pre-distortion coefficients for each operating frequency. Fortunately, it is possible to generate these coefficients automatically by means of a DSP "unpublished" [8]. Once a full group of coefficients is evaluated, they can be stored in a look-up table, and applied by the DSP to the input signal as necessary.

3 Transmitter prototype

The prototype constructed for the purposes of the project operated at a centre frequency of 220 MHz with a hopping bandwidth of 10 MHz, and a linearisation bandwidth of 5 kHz. To avoid instability caused by the second system poles, relatively wideband (compared with the bandwidth of the loop filters) op-amps with a unity-gain bandwidth of 10 MHz were used (the bandwidth of the loop filters was about 2 kHz). Wideband RF components with bandwidths of a few tens of MHz were chosen to cover all possible carrier frequencies. The RF amplifier was a class AB amplifier whose frequency response and non-linear characteristics at various carrier frequencies are shown in Figures 2 and 3 respectively. The required control signals were generated by means of a DSP program.

3.1 Transmitter transient response

The speed at which a transmitter responds to switching conditions may be assessed by measuring its transient response. In the case of a *switched-carrier* Cartesian transmitter, two forms of transient response measurement need to be performed:

- baseband transient response, which represents the transmitter speed of response to the switching conditions in its input modulation signal
- RF transient response, which is indicative of the transmitter speed of response to a switched carrier (i.e. a carrier which is switched off between consecutive hops)

In this section, the nature of both forms of transient response is investigated, and the results obtained from the hardware prototype are presented.

Baseband transient response

The baseband transient response of a Cartesian transmitter is determined by the size of its gain-bandwidth-delay (GBWD) product. Larger values of this parameter correspond to faster responses. Overshoots may be observed on the rising and/or falling edges of the output waveform when the GBWD product is too large [7]. The height of the overshoots may be reduced by lowering the GBWD product at the expense of the transmitter linearisation performance.

The baseband transient response of the transmitter prototype is shown in Figure 4. It can be seen that the output signal envelope takes about 6 μs to reach its maximum when the input signal envelope rises to its maximum in 4.2 μs: a fast response considering the narrowband nature of the transmitter input signals.

RF transient response

The DC signal introduced in the feedback path by the down-converter mixers is amplified by the loop gain, generated by the modules situated between the down-converter and the transmitter output. In steady state conditions, the effect of the mixer generated DC offset may be removed by pre-biasing the transmitter input signal. However, if the local oscillator is turned on and off, as in the case of a switched-carrier FH transmitter, the initial level of the DC offset can be large, and its effect severe when magnified by the loop action. The left hand plot of Figure 5 shows the output signal when the transmitter input was reduced to zero, and the 220 MHz carrier turned on and off. Although the carrier frequency was not changed in obtaining these results, the plot clearly demonstrates the transmitter RF transient response (the transmitter behaviour would be similar if the carrier frequency were changed).

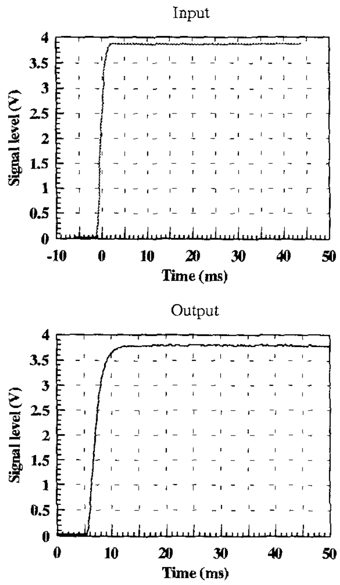


Figure 4: *Baseband transient response of the prototype*

The down-converter mixers react to the switched carrier by generating impulses which are amplified by the loop action. The presence of these overshoots can cause significant broadband interference, which is not acceptable, especially in a mobile environment. A simple technique for overcoming the problem of overshoots involves the reduction of the loop gain at the carrier signal switching instants. This can be done either at RF (i.e. reducing the RF amplifier and/or pre-amplifier gain) or baseband (i.e. reducing the gain of the loop baseband modules). To improve the transient response of the transmitter prototype, the loop gain was reduced by changing the gain of its first-order filters. The block diagram of the first-order filter whose transfer function is given by (2) (R is the parallel equivalent of r and R_f) is shown in Figure 6.

$$\frac{V_{out}}{V_{in}} = \frac{-R}{R_{in}(1 + RC\omega j)} \qquad (2)$$

By choosing r to be much smaller than R_f, the gain of the filters is reduced by a factor of $r/(r+R_f)$ when switch S (a CMOS switch) is closed. The filters bandwidth, on the other hand, is increased by the same factor, keeping the loop GBWD product constant. The large value of filter bandwidth results in much shorter transients which is another useful property of this technique.

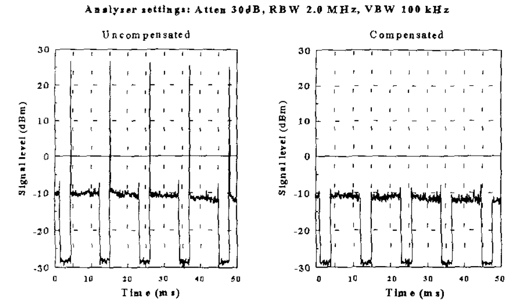


Figure 5: *Transmitter RF transient response (before and after applying gain control)*

The effectiveness of this technique when applied to the transmitter prototype is illustrated by the right hand plot of Figure 5. It can clearly be seen that the reduction in the loop baseband gain has decreased the height of the overshoots by about 35 dB, virtually removing the problem of RF transients. The control signal activating the switches was provided by the DSP, as discussed in the following section.

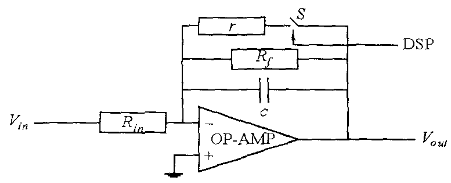


Figure 6: *Variable gain first-order filter*

3.2 Control signals

The transmitter control signals need to be carefully generated for successful performance. Such signals may be produced by means of a DSP. Some of the main functions of the DSP are:

- triggering frequency synthesizer
- controlling loop gain
- generation of the phase-shifter control signal
- input signal pre-distortion
- baseband signal shaping

The hopping carrier used with the transmitter prototype was generated by an HP 8645A Agile Signal Generator operated in its "external trigger - internal dwell" mode. In this mode, the carrier frequencies and the order in which they were applied were initially saved. The DSP was then used to generate a two-level signal (S_1 in Figure 7) which triggered the generator to turn on the carrier (external trigger). The time during which the carrier signal remained on was determined internally (internal dwell) by saving the required dwell time and hop rate prior to starting the hop sequence. With this generator, the carrier envelope started to rise about 12 μs (T_{idm}) after the generator was triggered, imposing a minimum limit on the value of the carrier "idle time" (T_{id}), i.e. the time during which the carrier remained off between any two consecutive hops. The total of the carrier rise and fall times $(T_{rs} + T_{fl})$ was about 3 μs.

The trigger signal, S_1 in Figure 7, was also employed to activate the CMOS switches whose function was discussed in the previous section. The undesirable overshoots resulting from the transmitter RF transient response appear as soon as the carrier signal is turned on, i.e. at instants such as t_1 and t_8. However, if the CMOS switches are already activated prior to the occurrence of the overshoots, the height of the overshoots may be reduced significantly. To ensure this effect, the duration of the trigger pulse, T_s, was chosen to extend just beyond the generator minimum idle time, T_{idm}. The right hand plot of Figure 5 shows the transmitter RF transient response, with a signal similar to S_1 activating both the generator and the switches.

To ensure transmitter stability, changes in the phase-shifter control voltage must occur while the loop gain is sufficiently low. The best time for changing this voltage is when the carrier signal is turned off, i.e. during the generator idle time (T_{id}).

As can be seen in Figure 7, the total hopping period can be calculated from

$$T_{hop} = T_{dwl} + (T_{rs} + T_{fl} + T_{id}) \tag{3}$$

Assuming a negligible trigger pulse duration compared with $(T_{rs} + T_{fl} + T_{id})$, the maximum hop rate which can be tolerated by the transmitter will be mainly determined by the speed of the frequency synthesizer, transmitter linearisation bandwidth, and in the case of digital transmission, number of symbols per hop and symbol pulse shape.

If the acceptable ratio of $(T_{rs} + T_{fl} + T_{id})$ to the total hopping period is given by n, then for a hopping frequency of m hops per second the following relationship must be valid.

When transmitting digital data (assuming a square pulse shape), the following expressions give m and n in terms of the total switching period T_{sw}, number of symbols per hop q, and the signal RF bandwidth B_{RF} (a function of transmitter linearisation bandwidth).

$$\frac{n}{m} = \overbrace{T_{rs} + T_{fl} + T_{id}}^{T_{sw}} \qquad n < 1 \tag{4}$$

$$n = \frac{T_{sw} \cdot B_{RF}}{2q + T_{sw} \cdot B_{RF}} \qquad m = \frac{B_{RF}}{2q + T_{sw} \cdot B_{RF}} \tag{5}$$

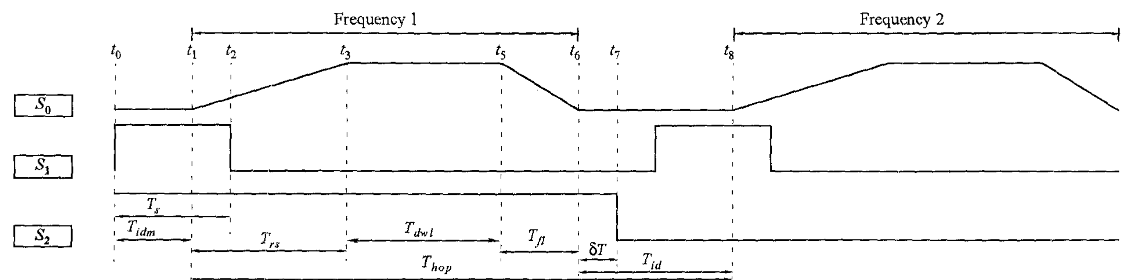
In the case of a 5 kHz transmitter, with $T_{sw} = 15\,\mu s$ and $q = 1$ symbol/hop, n and m are 3.6% and 2410 hops/s respectively. (5) shows that the hop rate increases with an increase in B_{RF} and/or a decrease in q and/or T_{sw}. As the RF bandwidth of the modulating signal is affected by the symbol pulse shape, the hop rate is also a function of the pulse shape. If raised cosine filters are used to shape the symbols, the above relationships will be replaced by those given by (6).

$$n = \frac{T_{sw} \cdot B_{RF}}{(\alpha + 1)\,(q + 1) + T_{sw} \cdot B_{RF}} \qquad m = \frac{B_{RF}}{(\alpha + 1)\,(q + 1) + T_{sw} \cdot B_{RF}} \tag{6}$$

where α is the filter roll-off factor. Keeping T_{sw}, B_{RF}, and q as above, the corresponding values of n and m for $\alpha = 0.5$ are evaluated as 2.4% and 1626 hops/s respectively.

3.3 Transmitter output spectrum

As the RF amplifier non-linear characteristics remained unchanged throughout the hopping bandwidth (Figure 3), the transmitter level of linearity was expected to stay the same at all frequencies of interest. The measured two-tone test response of the hopping transmitter (Figure 8) indicates the same level of linearity at various frequencies of operation, proving the validity of the theoretical prediction. The output spectrum of the hopping transmitter for carrier frequencies of 216, 217, 218, and 219 MHz



Legend

T_{idm} :Local oscillator minimum idle time
T_{dwl} :Local oscillator dwell time
T_{rs} :Local oscillator envelope rise time
S_1 :Frequency synthesizer and filter switch control signal (synthesizer in external trigger, internal dwell mode); logic 1 or 0
T_s :Trigger signal on period
S_0 :Local oscillator envelope
T_{id} :Total local oscillator signal idle time
δT :Delay between the bottom of local oscillator envelope ramp-down and phase-shifter voltage transition instant
T_{fl} :Local oscillator envelope fall time
S_2 :Phase-shifter control signal; analogue
T_{hop} :Hop period

Figure 7: *Transmitter DSP generated control signals and local oscillator envelope*

is shown in Figure 9, indicating no signs of instability. At the time of these measurements the hop rate was set at 220 hops/s. It could, however, be increased to a few thousand, depending on the parameters outlined in the previous section.

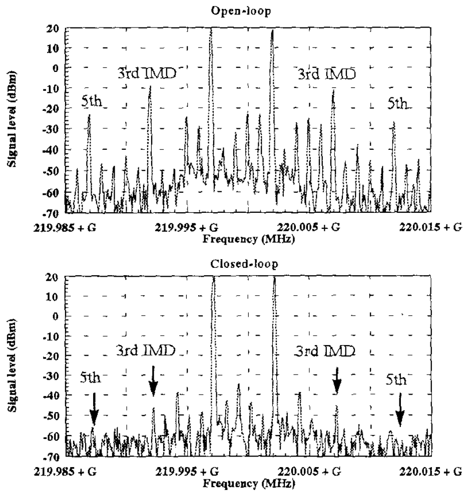


Figure 8: *Transmitter open- and closed-loop output spectra,* $G = -5, \ldots, -1, 0, 1, \ldots, 5$

4 Summary and conclusions

The results from a FH Cartesian transmitter prototype, constructed at Bristol, indicate that as long as the RF amplifier gain response and non-linear characteristics do not vary significantly with frequency, and its phase response is linear within the hopping bandwidth, it is perfectly possible to maintain the high level of linearity generated by the basic transmitter throughout a wide hopping bandwidth.

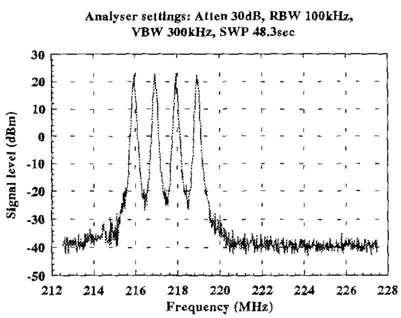


Figure 9: *Transmitter hopping spectrum (hopping bandwidth of 4 MHz)*

The transmitter baseband transient response, governed by the size of its GBWD product, is mainly determined by the characteristics of the RF amplifier (i.e. its non-linear characteristics and group delay). Its RF transient response, on the other hand, depends on the size of the loop gain generated by the components situated between the down-converter mixers and the transmitter output, and the shape of the carrier envelope applied to the down-converter mixers. Hence, the RF transient response can be significantly improved by shaping the latter and/or reducing the loop gain at the carrier switching instants. In the case of the transmitter hardware prototype, this was achieved at baseband by changing the gain of loop filters.

The results obtained from the prototype suggest that the Cartesian technique can be successfully used when hopping at low rates (e.g. a few thousand hops per second) as long as a hopping carrier with fast switching times is available. Although the main focus of this paper has been on the properties of the Cartesian transmitter under frequency hopping conditions, it must be borne in mind that most of the points discussed here are equally applicable to a TDMA transmitter.

The FH Cartesian transmitter provides the possibility of using spectrally efficient linear modulation techniques with frequency hopping for use in third generation communication systems. Efficient use of the valuable frequency spectrum, low adjacent and cochannel interference, large system capacity, small size, low weight, and higher power-efficiency are amongst the very attractive properties of such a transmitter.

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