standard deviation (SD) of the SIR estimation error was calculated for the two algorithms and the results are presented in Fig. 3.

The superiority of the Kalman filter algorithm is obvious at all mobile speeds. It is noteworthy that the SD for the new algorithm does not exceed 0.75dB, while this is up to 1.2dB for Steele's algorithm at mobile speed 96km/h. The adaption of the new algorithm to different mobile speeds is demonstrated by the SD increasing only 0.3dB with mobile speed increases from 12 to 96km/h, while the SD of Steele's algorithm increases by 0.6dB.

Conclusions: A new adaptive algorithm for the SIR estimation at the base station of a CDMA cellular system has been proposed, which utilises Kalman filtering for channel estimation, and calculates the signal as well as the interference power. It adjusts its parameters according to the mobile station speed to provide good results even at high speeds. These results are superior to those of the conventional algorithm [1].

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Calibration scheme for LINC transmitter

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A simple calibration scheme for correction of the path imbalance in a LINC transmitter is presented. By properly exchanging two LINC vector components and controlling a downconversion loop, the gain and phase imbalance are characterised by the baseband DSP/SCS. The error effects of the imbalance are compensated by introducing a pre-distortion term. Simulation results demonstrate that this scheme is sufficient to suppress the out-of-band spectrum for wireless communications.

Introduction: Linear modulations such as QPSK and QAM are desired for mobile communications owing to their superior spectral efficiency, but they require linear microwave amplification. Among the proposed power amplifier linearisation techniques, linear amplification with nonlinear components (LINC) is a promising approach, which can achieve high linearity and high power efficiency simultaneously [1].

A major drawback of LINC is its extremely tight tolerance on the matching between two power amplifier branches to achieve acceptably small out-of-band rejection. Several attempts have been made to correct for the imbalance through some kind of feedback approach. A 'phase-only' method was proposed in [2]. A simplex search algorithm was proposed in [3] to correct both gain and phase imbalance. The correction of these errors relies on the measurement of the out-of-band emission, which in turn sets a lower limit on the calibration time of approximately 1 to 2s. A direct search method was proposed in [4] to correct the gain imbalance as well as the consequent phase imbalance from AM-PM conversion. We proposed a calibration scheme [5], in which the gain and phase imbalance are evaluated with a set of calibration signals. However, the application of this scheme is limited because the specific time slots are required for calibration before data transmission.

This Letter describes a new technique to calibrate the gain and phase imbalance in the background during normal transmission of the signal. The imbalance information is abstracted from the mixed product of LINC output and two signal components.

Principle of LINC: The basic principle of LINC is to represent an arbitrary bandpass signal $s(t) = a(t)e^{j\theta(t)}$, where $0 \le a(t) \le V_m$, by means of two out-phased and constant envelope signal components

$$S_1(t) = s(t) - e(t) \qquad S_2(t) = s(t) + e(t) \qquad (1)$$

where the quadrature signal $e(t) = js(t)\sqrt{[V_m^2/a^2(t) - 1]}$. These two signals are amplified separately and then combined. With the combining network, the in-phase signals add together while the quadrature signals cancel each other, hence an amplified replica of the input is obtained. In the event of any gain or phase mismatch between two amplifier branches, the cancellation of quadrature signals is incomplete, which leaves a residue in adjacent channels and introduces adjacent channel interference (ACI) [3].

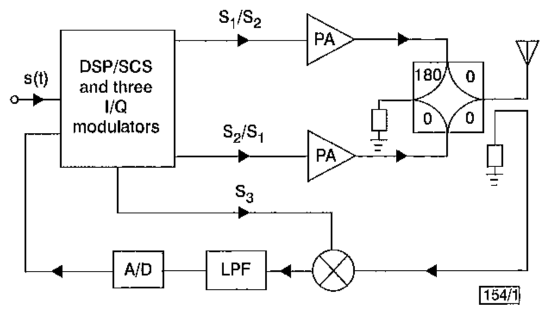

Fig. 1 *Simplified LINC diagram with calibration loop*

Calibration algorithm: The proposed scheme characterises the gain and phase imbalance through a feedback loop, as shown in Fig. 1. Note that the D/A converters, lowpass filters and I/Q modulators are combined with the DSP/SCS for simplicity. A small portion of the LINC output is withdrawn and multiplied by the signal S_3. This mixed signal is then lowpass filtered, D/A converted and sent back to the DSP which searches for the maximum and minimum signal values. The baseband digital circuit switches the branch S_3 to the upper or lower amplifier branch. For each case, the DSP regularly exchanges two LINC vectors S_1 and S_2 in two amplifier branches back and forth. As will be shown, this will guarantee that the DSP finds two maxima and two minima, and these four quantities completely determine the gain and phase imbalance. Assuming that G_0 is the amplifier gain and ϕ_0 is the path phase delay, and the gain and phase imbalance are Δg and $\Delta\phi$, respectively, the LINC output will be

$$\begin{aligned} S(t) = {} & G_0 V_m \cos[\omega_c t + \theta(t) \mp \psi(t) + \phi_0] \\ & + (1 + \Delta g) G_0 V_m \cos[\omega_c t + \theta(t) \pm \psi(t) + \phi_0 + \Delta\phi] \end{aligned} \qquad (2)$$

where $\psi(t) = \cos^{-1}[a(t)/V_m]$, ω_c is the carrier frequency, and '∓/±' are the consequence of exchanging two vectors S_1/S_2. When S_3 connects to the upper amplifier branch, after lowpass filtering we get

$$\begin{aligned} S_{LPF} = {} & \frac{1}{2} G_L V_m \cos\phi_L \\ & + \frac{1}{2}(1 + \Delta g) G_L V_m \cos[\pm 2\psi(t) + \phi_L + \Delta\phi] \end{aligned} \qquad (3)$$

where G_L is the loop gain and ϕ_L consists of the loop delay and the phase shift introduced by the downconversion mixer. This is a baseband-amplitude-determined sinusoidal signal offsetted by a DC constant. The DSP searches for its maximum/minimum value, depending on the sign of $2\psi(t)$. When this value is found, the DSP exchanges S_1/S_2 and determines the minimum/maximum value. Since $-180° \le \pm 2\psi(t) \le 180°$, the maximum and minimum signal values will be determined. Now two quantities are obtained by combining the maxima and minima:

$$S_{a+} = max + min = G_L V_m \cos\phi_L \qquad (4a)$$

$$S_{a-} = max - min = (1 + \Delta g) G_L V_m \qquad (4b)$$

Similarly, S_{b+} and S_{b-} can be computed while the DSP switches S_3 to the lower amplifier branch:

$$S_{b+} = (1 + \Delta g)G_L V_m \cos(\phi_L + \Delta\phi) \quad (5a)$$

$$S_{b-} = G_L V_m \quad (5b)$$

Compare eqns. 4 and 5, the gain imbalance is given by

$$\Delta g = \frac{S_{a-}}{S_{b-}} - 1 \quad (6)$$

The phase imbalance is resolved from $\cos(\phi_L + \Delta\phi)$ and $\cos\phi_L$, and the accuracy of estimation depends on ϕ_L, considering $\Delta\phi$ is a small quantity. The maximum resolution results if $\cos\phi_L = 0$, and the accuracy decreases half if ϕ_L varies ±60° with respect to the optimum point. The optimum condition can be achieved by introducing a proper phase shift in S_3 branch such that $S_{a+}/S_{b-} \simeq 0$. As a result, we obtain

$$\Delta\phi = \pm\left(\frac{S_{b+}}{S_{a-}} - \frac{S_{a+}}{S_{b-}}\right) \quad (7)$$

where '±' is determined by the sign of ϕ_L. Alternatively, the phase imbalance can also be estimated by eqn. 7 if ϕ_L is not well controlled. The estimation error would be < 10% for a ±25° variation on ϕ_L with respect to the optimum point. This error only scales the estimated result. Hence the correction of the phase imbalance can be iteratively accomplished, provided the proper allowable imbalance level is set.

Simulation results: We simulated this calibration scheme with Cadence SPW according to the configuration of Fig. 1. S_3 was initially connected to the upper amplifier branch. To determine the correct maxima/minima, the baseband signal amplitude needs to get below a certain low level before exchanging S_1/S_2, e.g. $a(t) < V_m/\sqrt{2}$ for the optimum condition. The DSP then switches S_3 to the lower amplifier branch and determines another maxima/minima. At the end of iteration, the path imbalance is calculated and compensated by scaling and phase shifting the lower amplifier branch signal. These steps are repeated if necessary. The exchanging of S_1/S_2 may present a ripple to the DSP owing to the possible phase discontinuity, which in turn affects the measured maxima/minima. This error effect can be minimised by exchanging S_1/S_2 only if these two vectors are close enough to each other, i.e. $a(t) \simeq V_m$, or by holding the DSP search algorithm until the LPF settles down after exchanging.

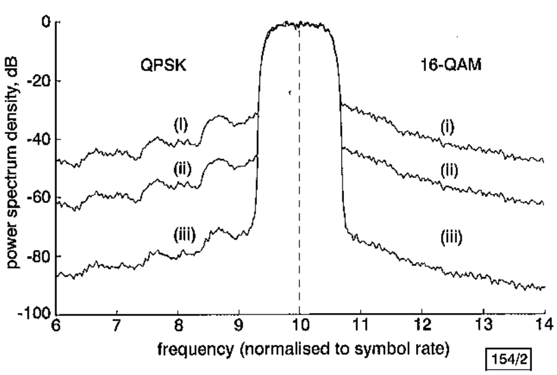

Fig. 2 *Simulated output spectra for QPSK/16-QAM modulated data*

(i) without correction (Δg = 1.0dB, $\Delta\phi$ = 7°)
(ii) without correction (Δg = 0.2dB, $\Delta\phi$ = 1.3°)
(iii) with correction

Fig. 2 shows the simulation results, where the modulation schemes used were QPSK (left) and 16-QAM (right), with square root raised cosine filtering of roll-off 0.35. The carrier frequency was ten times higher than the symbol rate and the sampling frequency was sixteen times higher than the carrier frequency. Because of the approximation in estimating the phase imbalance ($\phi_L \simeq 60°$ in our simulation), the calibration was accomplished iteratively and the algorithm converged in 2 to 3 iterations. With correction, the out-of-band spectrum is suppressed under −71dBc for QPSK and −73dBc for 16-QAM modulated data.

Conclusions: A novel calibration scheme for correction of the path imbalance in a LINC transmitter is presented. In this scheme, the baseband DSP regularly exchanges two LINC signal components and controls the downconversion loop, and the gain and phase imbalance are characterised iteratively. Simulation results demonstrate that this scheme is sufficient to suppress the out-of-band spectrum for wireless communications.

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Tapped delay line model for band-limited multipath channel in DS-CDMA mobile radio

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It is explained that a correlated tapped delay line model needs to be assumed for transmission performance estimation using Rake combining, even though the propagation channel itself can be characterised as a wide sense stationary uncorrelated scattering channel. Fading correlation between different tap gains can be computed by taking into account the transmit and receive chip pulse shaping filters. A simple example, showing how correlation affects performance estimation in Rake combining, is presented, assuming a two-path Rayleigh fading propagation channel.

Problem: In mobile wireless communications, owing to reflection by obstacles such as buildings, there are many propagation paths between a transmitter and a receiver. If the time delays of multipaths cannot be neglected, the propagation channel is called a frequency selective channel. Assuming a wide sense stationary uncorrelated scattering (WSSUS) channel, the equivalent lowpass impulse response of the propagation channel at time t, owing to an excitation at time $t - \tau$, can be expressed as [1]

$$h_c(t,\tau) = \sum_{m=0}^{\infty} \xi_{c,m}(t)\delta(\tau - \tau_{c,m}) \quad (1)$$

where $\xi_{c,m}(t)$ and $\tau_{c,m}$ are, respectively, the complex-valued random path gain and time delay of the mth path, $\xi_{c,m}(t)$ satisfies $\Sigma_{m=0}^{\infty} E[|\xi_{c,m}(t)|^2] = 1$ with $E[\cdot]$ denoting ensemble averaging, and $\delta(x)$ is the delta function. For the WSSUS channel assumption, $|\xi_{c,m}(t)|$ are statistically independent random processes.

When observing the propagation channel over a limited bandwidth, the band-limited channel is seen as a different propagation channel from the real one. For signal transmission covering the bandwidth of B Hz, the equivalent channel can be completely represented by a tapped delay line model [2]:

$$h(t,\tau) = \sum_{l=0}^{\infty} \xi_l(t)\delta\left(\tau - \frac{l}{B}\right) \quad (2)$$

This model is widely used for performance estimation of a direct sequence code division multiple access (DS-CDMA) system with Rake combining and a chip rate of $B = 1/T_c$. However, as we will show in the following, the random path gain $\xi_l(t)$ in eqn. 2 is a