where $r\theta$ is the dynamic quality factor of the varactor diode, r is the modulation coefficient of the varactor diode, and θ is the static quality factor of the single frequency of varactor

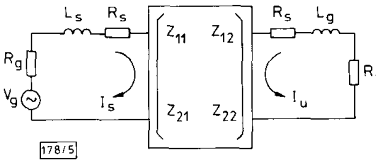

Fig. 5 *Equivalent circuit of upconvertor*

diodes. In eqns. 7 and 8, the coefficient 1/2 is used to consider the consistence of the two varactor diode properties because the signal circuits and sum-frequency circuit are connected in parallel.

The gain of the conversion power of the sum-frequency upconvertor is

$$G_{ut} = \frac{P_u}{P_{sa}} = \frac{I_u R_u}{V_g^2/4R_g} = \frac{4(\gamma\theta)^2 R_u R_g R_s^2}{[(R_g + R_s)(R_u + R_s) + \omega_s/\omega_u(\gamma\theta)^2 R_{su}^2]^2} \quad (9)$$

As $R_g = R_s = R_m = R_s/\omega_s/\omega_u(\gamma\theta)^2 + 1$, the maximum gain of the conversion power is

$$G_{utm} = \frac{(\gamma\theta)^2}{[\sqrt{(\omega_s)/\omega_u(\gamma\theta)^2 + 1} + 1]^2} = \frac{(\gamma\theta)^2 R_s^2}{(R_m + R_s)^2} \quad (10)$$

After determination of varactor diodes and their quality factors, the upconvertor can be designed in accordance with eqns. 7–10.

Test results:
(i) The optimum pumping power is ~60 mW.
(ii) The maximum power output of the sum-frequency signal is 1·34 mW.
(iii) The conversion loss of frequency is 2·98 dB.
(iv) The bandwith of 3 dB is from 630 to 1000 MHz.
(v) 2·06 and 2·09 VSWR have been achieved at the signal and pumping ports, respectively.
(vi) The isolation between ports is greater than 40 dB.
(vii) Image control is greater than 40 dB.
(viii) The purity of spectrum of the sum-frequency signal is as follows: the width of spectrum of the input signal and sum-frequency signal are both 2 kHz; there is no parasitic disturbing frequency spectrum observed even when the amplitude of the sum-frequency signal reaches 30 dB.
(ix) Phase-to-phase characteristics: phase variation of 180° of the input signal results in a 180° phase variation of output signal; the input phase at intervals of 10° results in a phase variation of output signal within 10° ± 1°.
(x) Amplitude-to-phase characteristics: an amplitude variation of 23 dB of input signal results in a phase variation of output signal within 0°–4°.
(xi) Phase-to-amplitude characteristics: phase variation of 720° of input signal causes a variation of ±0·1 dB of output power.
(xii) Amplitude-to-amplitude characteristics: a variation of 33 dB of input signal causes 33·08 dB variation of output power the difference between them is 0·08 dB.

Conclusion: The test results show that the new circuit designed has an excellent performance. It has a high fidelity of amplitude and phase, and an output signal frequency spectrum of high purity in its linear operating frequency band.

A new circuit to realise the conversion of microwave signals to millimetre-wave signals with a high fidelity is proposed. This offers a useful and economic way to make full use of the millimetre-wave band.

DENG SHAOFAN
WU QUN
QIU JINGHUI

Harbin Institute of Technology
Harbin, (150006), People's Republic of China

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OPTIMUM RATE REED SOLOMON CODES FOR FFH/CDMA MOBILE RADIOS

Indexing terms: Codes and coding, Mobile radio systems, Communication networks

The forward link (base to mobile) user capacity of Reed Solomon (RS) coded FFH/CDMA channels is investigated. A new error-and-erasure correction scheme for the RS coded FFH/CDMA channels is proposed. It is shown that the proposed error-and-erasure correction scheme for the RS(255, 223) code with the optimal code rate 0·87 can increase the normalised user capacity by 17% over that without coding when the deletion probability is 0·01.

Introduction: In the fast frequency hopping code division multiple access (FFH/CDMA) system proposed by Goodman *et al.*, the input symbol (= K bits) to be transmitted is divided in time into several chips, and a random symbol sequence whose rate is equal to the chip rate is added as an address to each chip with modulo-2^K.[1] One of 2^K frequencies, corresponding to one chip added by the random address, is transmitted over the chip interval. Energy detection is used for FFH signal reception. The same frequency band is shared by many users, and the transmitted symbol is extracted from the detection matrix indicating detected frequencies.

In the analysis of Reference 1, the received symbol is regarded as an error if the dehopped detection matrix has more than one row with the greatest number of entries. It was shown in Reference 1 that when the deletion probability is 0·01, 183 forward link (base to mobile) users transmitting 32 kbit/s speech (or data) with a bit error rate (BER) of less than or equal to 10^{-3} can be accommodated in a 20 MHz bandwidth. One approach to increase the user capacity was proposed in Reference 2 and used address coding. Another powerful strategy is to use coding for error protection. Kawahara and Matsumoto* investigated the theoretical limit of user capacity using the channel cutoff rate of the FFH/CDMA channel, and with deletion-free transmission, the maximum user capacity should be realised with rate 0·85 codes.

This Letter investigates the forward link user capacity of FFH/CDMA mobile radio channels with Reed Solomon (RS) codes. A new error-and-erasure correction scheme is proposed in which the received symbol is regarded as an erasure if there is more than one row having the greatest number of entries in the dehopped detection matrix. Decoding performance of the RS codes with the proposed error-and-erasure correction is theoretically analysed. The user capacity of a coded FFH/CDMA channel is then evaluated.

System Model: Fig. 1 shows the system model analysed. The input bit stream to be transmitted is stored in a K bit shift register and segmented into K bit symbols. The K bit symbol

* KAWAHARA, T., and MATSUMOTO, T.: 'User capacity limit of a coded FFH/CDMA mobile radio channel', unpublished

sequence is then encoded into the coded symbol sequence of an RS(n, k) code defined over the Galois field of GF(2^K). The

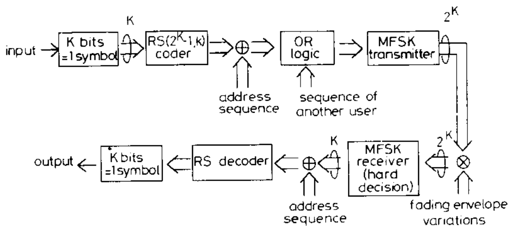

Fig. 1 *System model*

coded symbol with symbol duration of T is then divided in time into L chips, each of which has chip duration of T/L, and a random address sequence with rate of L/T is added to the chip sequence with modulo-2^K. The output chip sequence is fed to an OR logic circuit to 'accommodate' other user chips, and then one of the 2^K frequencies corresponding to one chip is transmitted over the chip interval.

In the receiver, received power of each of the 2^K frequencies is detected by an energy detector. Hard decision is used for detection. Both the transmitter and receiver know the predetermined address sequence, and the address sequence is then added with modulo-2^K to the detection matrix indicating received detected frequencies. The detection matrix is the input to the decoder of the RS code. A new method of identifying erasure symbols for an RS coded FFH/CDMA channel is proposed. The transmitted symbol is then estimated using the error-and-erasure correction scheme.

Error-and-erasure correction for decoding of RS code: The received symbol is regarded as an erasure if there is more than one row with the largest number of entries in the dehopped detection matrix. Fig. 2*a* shows an example of the dehopped detection matrix indicating symbol erasure encountered in the

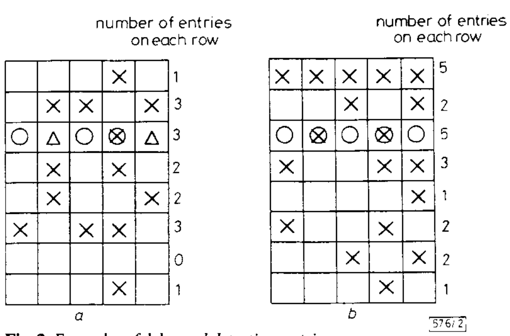

Fig. 2 *Examples of dehopped detection matrix*

a Detection matrix with three incomplete rows containing greatest number of entries
b Detection matrix with two complete rows
○ transmitted desired frequency
× insertion due to interference
△ deleted chip

FFH/CDMA channel with $L = 5$ and $2^K = 8$. The detection matrix has three incomplete rows containing the greatest number of entries, and thus the rows are regarded as indicating a symbol error in the system of Reference 1. Fig. 2*b* shows another example of the dehopped detection matrix indicating symbol erasure for $L = 5$ and $2^K = 8$. There are two complete rows, and thus, this matrix outputs a symbol error in the system of Reference 1.

The error and erasure correcting capability of the RS code is

$$2s + e + 1 \le d \tag{1}$$

where e and s are the number of symbol errors and the number of the erasures, respectively, produced in the dehopped detection matrix, and d is the minimum distance of the RS code. The user capacity of the RS coded FFH/CDMA channel is, thus, increased with the erasure-and-error correction capability over that possible with error correction alone.

Performance analysis: The formulas necessary to analyse the decoding performance of an RS coded FFH/CDMA channel are shown in Table 1 of Reference 1; these formulas will be used in this Letter without giving derivations. The average probability p_I of insertion due to interference in the dehopping process is given by

$$p_I = [1 - (1 - 2^{-K})^{M-1}](1 - p_D) \tag{2}$$

where p_D is the average deletion probability that the transmitted frequency is not detected by the energy detector and M is the number of users sharing the same FFH/CDMA channel. p_D is expressed as

$$p_D = \exp\left\{\frac{-\beta^2}{2(1+\Gamma)}\right\} \tag{3}$$

where Γ is the average received signal to noise power ratio (SNR), and β is the threshold normalised with respect to the receiver noise power.[3] Another factor that affects the value of p_I is false alarm probability p_F, which is expressed as a function of β and was taken into account in Reference 1. However, because for large Γ values p_D is approximately expressed as a function of β^2/Γ, p_D can be set any value independently of p_F. Thus we assume $p_F = 0$.

The dehopped detection matrix outputs an erasure symbol with probability p_E given by

$$p_E = \sum_{i=0}^{L}\left(P_C(i)\sum_{k=1}^{2^K-1}P(i,k)\right) + \sum_{i=0}^{L}\left(P_C(i)\sum_{j=i+1}^{L}\sum_{k=2}^{2^K-1}P(j,k)\right) \tag{4}$$

where $P_C(i)$ is the probability that the correct row in the detection matrix has i entries, and $P(n, k)$ is the probability that there are k undesired rows in the detection matrix and the greatest number of entries in the k undesired rows is n (see Reference 1). The symbol error probability p_W is expressed as

$$p_W = 1 - \sum_{i=0}^{L} P_C(i)P(i, 0) \tag{5}$$

A decoding error results if eqn. 1 is not satisfied. Therefore, assuming that insertion and deletion occur randomly, the code word error probability (WER) is expressed as

$$WER = 1 - \sum_{e=0}^{d-1}\sum_{s=0}^{t}\binom{N}{e}\binom{N-e}{s} \times p_E^e(p_w - p_E)^s(1 - p_W)^{N-e-s} \tag{6}$$

where $t = \lfloor (d - e - 1)/2 \rfloor$ and $\lfloor X \rfloor$ denotes the largest integer less than or equal to X. Using eqn. 6, the WERs were numerically calculated for RS(255, k) codes defined over GF(2^8) with $L = 19$ and $K = 8$, where k = 153–253. The number $Mk/255$ of the users, whose $WER \le 2 \times 10^{-3}$ (this roughly corresponds to the bit error rate of 10^{-3}), normalised by the code rate $k/255$ was then evaluated for each value of M. Fig. 3 shows the calculated normalised user numbers against code rate $k/255$ for both error correction alone and the proposed erasure-and-error correction scheme for the RS codes. The optimal code rate is 0·87 at which a coded FFH/CDMA channel with $L = 19$ and $K = 8$ can accommodate 208 users with error correction when $p_D = 0{\cdot}01$. This user capacity can further be increased to 213 users by the proposed error-and-erasure correction scheme, which is an increase of 17% over that without coding.[1] Also, the user capacity evaluated from the channel cutoff rate* achieved by hard decision maximum likelihood decoding is plotted to show that the user capacity should be maximised by rate 0·85 codes. This is almost the

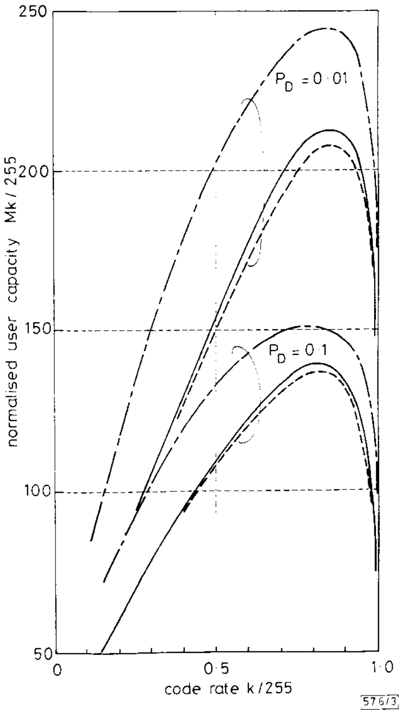

Fig. 3 *Normalised user capacity*

FFH/CDMA channel
RS(255, k) code
– – – – error correction
——— erasure-and-error correction
— — — maximum likelihood decoding

same as the optimal rate of the RS codes (rate 0·87 RS(255, 223) code). Furthermore, the normalised user capacity of 213 with the proposed erasure-and-error correction scheme is about 87% of the capacity limit with hard decision maximum likelihood decoding (capacity limit is 244 users).

Conclusion: The forward link user capacity of Reed Solomon coded FFH/CDMA channels was investigated. A new error-and-erasure correction scheme for an RS coded FFH/CDMA channel was proposed. The received symbol is regarded as an erasure if there is more than one row with the largest number of entries in the dehopped detection matrix. It has been shown that with the proposed error-and-erasure correction, the user capacity normalised by the code rate is increased by 17% over that without coding, when the deletion probability is 0·01.

T. KAWAHARA
T. MATSUMOTO

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NTT Radio Communication Systems Laboratories
1-2356, Take, Yokosukashi, Kanagawa-ken, 238, Japan

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LOW-LOSS LEAD FLUORIDE-COATED SQUARE WAVEGUIDES FOR CO_2 LASER LIGHT TRANSMISSION

Indexing terms: Waveguides, Lasers

Small-core (1 mm × 1 mm) phosphor bronze square waveguides whose inner two or four walls are coated with a thin lead fluoride layer have been fabricated by using vacuum-evaporation and assembly techniques. Straight waveguide transmissions as high as 80% are obtained even when the waveguides are bent with a bending radius of 50 cm in 1 m long waveguides.

Introduction: Realisation of low-loss waveguides with small bending losses is of great concern for uses of high powered CO_2 lasers.[1–6] In a recent report, it was shown that 1 kW-class CW CO_2 laser light was transmitted by a germanium-coated silver hollow waveguide 1·5 mm in diameter.[7] An output power of 2·6 kW of CW CO_2 laser light was also transmitted through a 2 m long waveguide 1·7 mm in diameter for an input power of 3 kW.* Therefore, transmission of kilowatt-class CO_2 laser light is possible by using small-core dielectric-coated waveguides. To increase the power capability and to improve the beam quality from the waveguides, it is essential to reduce the waveguide losses in the small-core waveguides.[8]

In this Letter, a simple method is employed to fabricate low-loss dielectric-coated waveguides with a square cross-section. 1 m-long lead fluoride (PbF_2)-coated waveguides with cross-section 1 mm × 1 mm are fabricated by using vacuum-evaporation and assembly techniques. Transmission losses in straight and bent waveguides are measured by the coherent CO_2 laser light and the method is found to be very useful for fabricating very low-loss waveguides.

Design of dielectric-coated square waveguides: Consider two types of dielectric-coated metallic square waveguide as shown in Fig. 1, where one (*a*) or two (*b*) facing metallic walls are properly coated with a thin dielectric layer. Let $n_0(\simeq 1)$, $a_1 n_0$, and k_0 be the refractive index of hollow core, refractive index of dielectric layer, and the wavenumber in free space, respectively. When the thickness of dielectric layer d is

$$d = \frac{\pi}{2n_0 k_0 (a_1^2 - 1)^{1/2}}$$

the E_{11}^y mode has a minimum attenuation constant and all modes with x polarisation are highly attenuated in the small-core waveguides of Fig. 1*a*.[9] On the other hand, in the waveguide of Fig. 1*b*, either the E_{11}^x or E_{11}^y mode and the circularly polarised mode have minimum loss when thickness d satisfies

$$d = \frac{1}{n_0 k_0 (a_1^2 - 1)^{1/2}} \tan^{-1}\left(\frac{a_1}{(a_1^2 - 1)^{1/4}}\right)$$

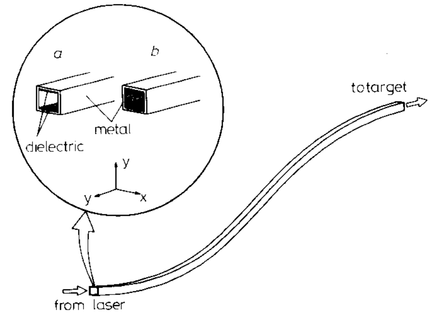

Fig. 1 *Waveguide structures*

a Two wall coated square waveguide
b Four wall coated square waveguide

* HONGO, A.: Private communication