the next frame at equilibrium, divided by the expected number of talkspurts in the current frame at equilibrium.[3] Under equilibrium conditions, the expected number of talkspurts which will not persist is the sum of the expected new arrivals and the expected departures of talkspurts in a frame. Thus

$$CF = \frac{\frac{\lambda}{\mu} - 2\lambda T}{\frac{\lambda}{\mu}} \tag{2}$$

Since the mean talkspurt length $(1/\mu)$ is generally much greater than twice the frame time $(2T)$, it is seen that the CF is close to unity. λ is the mean total talkspurt arrival rate.

Hence, to reduce the VT update time, larger VT clock speeds, η, could be used with variable length packets, since collisions between existing users never occur if $\eta < T_s/\tau$. Although a large clock speed will increase the number of collisions between new and existing users, existing users no longer discard their packets in the event of a collision, but transmit immediately after the jamming signal. At high load, packet lengths will be greater than P_s, and so packet separation times on the VT axis will be greater than T_s. The ratio of the number of effective reservation slots to new users is still adequate, if $\eta < T_s/\tau$ because of the high correlation factor.

This adaptive R-VT-CSMA (AR-VT-CSMA) protocol achieves a unique reservation time on the VT axis for existing users, which depends upon traffic intensity. It can improve throughput by using a larger VT clock speed than R-VT-CSMA, since the existing user reservation times are separated by a packet transmission time.

Simulation results: To examine the performance of the synchronous version of AR-VT-CSMA, a simulation was written in SLAM with the following assumptions:

* The distribution of talkspurts and silences at a voice station was based upon Brady's results.[4] The mean talkspurt and silence were 1·34 and 1·67 seconds, respectively.

* A voice station only transmitted packets during a talkspurt. The voice encoding rate (V) was 64 kbit/s, with 80 bit voice packet headers.

R-VT-CSMA/CD: 20 ms frame time, $\eta = 7$.

AR-VT-CSMA/CD: $P_{min}/V = 20$ ms, $P_{max}/V = 40$ ms, $\eta < T_s/\tau$.

* Data packets had a Poissonion arrival distribution from three network sources.

2000 bit data packets were used, with $\eta = 7$.

* The capacity of the LAN was 2 Mbit/s, with an end-to-end propagation delay of 10 μs.

Fig. 1 shows the voice bit loss as a function of the number of voice stations for a voice only load using R-VT-CSMA and AR-VT-CDMA/CD. Fig. 2 shows the data packet delay for various numbers of voice stations and with a constant network data load of 30%. It can be seen that AR-VT-CSMA

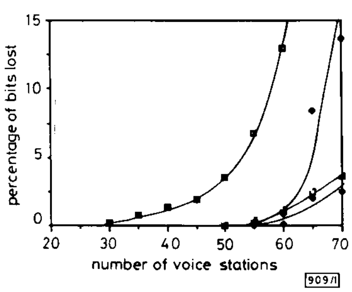

Fig. 1 *Bits lost at various dock speeds for voice only load*

□ R-VT-CSMA $\eta = 7$
◆ AR-VT-CSMA $\eta = 10$
◆ AR-VT-CSMA $\eta = 10$
□ AR-VT-CSMA $\eta = 40$
◇ AR-VT-CSMA $\eta = 65$

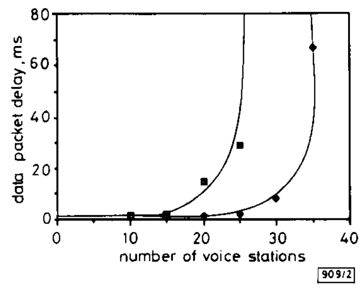

Fig. 2 *Data packet delay*

30% data load
□ R-VT-CSMA $\eta = 7$
◆ AR-VT-CSMA $\eta = 65$

can support many more voice stations and give reduced data packet delay over R-VT-CSMA. This improvement is due to a reduction in the voice VT clock update interval, existing users transmitting after collisions and increased voice packet header efficiency for the same maximum voice delay of R-VT-CSMA.

Conclusions: In this letter, an extension to the voice/data MAC protocol R-VT-CSMA has been presented. It has been shown that adaptive R-VT-CSMA can support more voice stations and give a lower data packet delay than R-VT-CSMA by increasing the VT clock speed, η. Voice packet reservation is maintained through the use of variable length packets.

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COMMENT

SLOTTED ALOHA RADIO NETWORKS WITH PSK MODULATION IN RAYLEIGH FADING CHANNELS

A recent letter proposed analytical results for receiver capture in ALOHA radio networks, considering the effects of binary PSK modulation, noise and packet size. A number of improvements are offered in this comment.

Introduction: The decision variable for synchronous bit extraction from a test signal with amplitude α_0 in the presence of N contending interferers is[1]

$$v = \alpha_0 a_0 + \sum_{j=1}^{N} \alpha_j a_j \cos\theta_j + n_i \tag{1}$$

where $a_j (= \pm 1)$ represents the bindary phase modulation of the jth carrier. In a Rayleigh fading channel, the in-phase components $\alpha_j \cos\theta_j$ of the N interferers and the additive noise n_i are independent Gaussian variables. With fast Rayleigh fading, the interference terms in eqn. 1 are statistically independent from bit to bit. In this event, the joint interference signal may be added to the Gaussian noise. The long-term mean joint interference power offered to the synchronous detector is

$$\bar{P}_N = \tfrac{1}{2}\sigma_i^2 + \tfrac{1}{2}\sigma_q^2 = \sum_{j=1}^{N} r_j^{-4} \tag{2}$$

σ_i^2 and σ_q^2 are the variances of the joint in-phase component (as in eqn. 1) and the joint quadrature component, respectively. r_j is the normalised distance between the jth terminal and the central receiver. A perfect synchronous detector sees v with long-term mean interference plus noise power $\bar{p}_N + N_0$. The corresponding conditional bit error permeability is

$$p_b(\text{error} \mid p_0, \bar{p}_N) = \frac{1}{2}\operatorname{erfc}\left\{\sqrt{\left(\frac{p_0}{\bar{p}_N + N_0}\right)}\right\} \tag{3}$$

given the power of the test signal $p_0 = \frac{1}{2}\alpha_0^2$, which is exponentially distributed because of Rayleigh fading of the amplitude α_0.

Comments:

(A) In contrast to eqn. 3, the instantaneous power p_N of the joint interference sample

$$p_N = \frac{1}{2}\left(\sum_{j=1}^{N} \alpha_j a_j \cos\theta_j\right)^2 + \frac{1}{2}\left(\sum_{j=1}^{N} \alpha_j a_j \sin\theta_j\right)^2 \tag{4}$$

was indiscriminately added to the average noise power N_0 in Reference 1.

(B) In Reference 1, a probability of bit error $P_{bn}(\bar{p}_0)$, spatially averaged for interference, was calculated by integrating over all values of $\bar{p}_N$, and subsequently used for calculation of a packet error probability. This sequence of statistical operations assumes that each of the N interfering terminals moves to a new, entirely uncorrelated position $\bar{r}_j$ after each bit in the packet. This appears inconsistent with selecting a test packet having fixed mean power $\bar{p}_0$ during all L bits of the packet,[1] and also with normal rates of mobility of contending interferers.

(C) For slow fading, the packet error approximated was by a bit error. This approximation is poor for a test signal buried in Gaussian noise and interference, since the bit error rate and the packet error rate for the test packet in this case are limited by

$$\text{BER} \to \tfrac{1}{2}$$
$$\text{PER} \to 1 - L/2 \to 1 \qquad \text{for large } L \tag{5}$$

(D) The possibility of correct reception of any contending packet other than the (*a priori* selected) test packet was not considered in Reference 1. Even if the receiver actually locks to this test packet, events with $\alpha_j \cos\theta_j \gg \alpha_0$ (e.g. if $r_j \ll r_0$) may still contribute to the total throughput, despite a synchronisation error θ_j.

Comparison: With regard to comments A and B, the throughput S_c for the selected test packet in fast fading becomes

$$S_c = \sum_{n=0}^{\infty} \frac{G^{N+1}e^{-G}}{(N+1)!} \int_0^{\infty}\int_0^{\infty} 4Nrs \exp\left\{\frac{-\pi}{4}(r^4 + N^2 s^4)\right\} \times \left[\frac{1}{2} + \frac{1}{2}\sqrt{\left(\frac{r^{-4}}{r^{-4} + s^{-4} + N_0}\right)}\right]^L dr\, ds \tag{7}$$

which is shown in Fig. 1 for $L = 10$ and $L = 50$. Slow fading (C) is believed to require more elaborate simulation models,

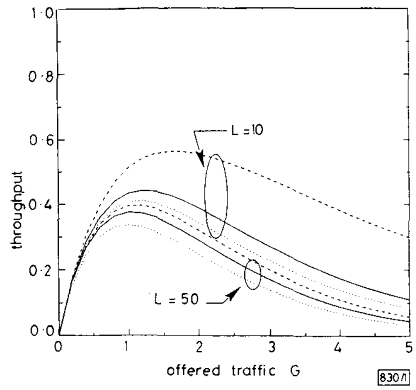

Fig. 1 *Total throughput against traffic*
BPSK modulation
SNR = 30 dB
- - - - proposed method
······ Reference 1
——— including comments A and B

e.g., Reference 2. The impact of selecting the favourite test packet (D) randomly is approximated (ignoring θ_i) by multiplying the capture probability in eqns. 6 and 7 in Reference 1 by $N + 1$, as in standard papers on the overall throughput S_t of slotted ALOHA with capture.[3]

Acknowledgment: The authors are grateful for discussions with Prof. K. Pahlavan and Prof. J. C. Arnbak on an earlier version of this comment.

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REPLY

We agree that any mathematical problem can be solved by different assumptions. Some assumptions are more realistic than others. We had made new sets of more practical assumptions for the same problem which were submitted for publication. The new work[2] includes an exact solution and two bounds for the problem. The relevance of the above comments is also discussed.

As far as the new eqn. 7 and the new plot in the revised comments are concerned; we used comment D and eqn. 7 to determine S_t for smaller packet lengths of ($L = 1$ and $L = 5$) which is shown in Fig. 1 (our plots are using more terms in the summation, so the tails of the curves are more straight than those sketched in their graph). For both cases ($L = 1$ and $L = 5$) the throughput is greater than one for high values of attempted traffic. This is unacceptable since a receiver can not receive more than one packet per slot under any circumstances. This observation raised some serious questions on the validity of assumption D and the qn. 7.

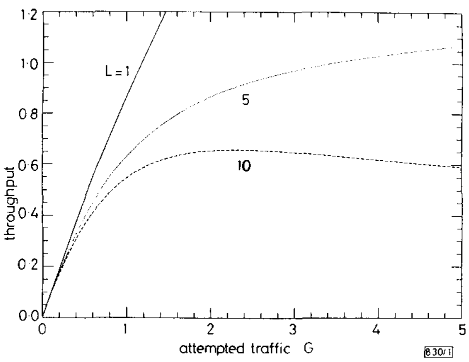

Fig. 1 *Total throughput against traffic for various packet sizes*
BPSK modulation
SNR = 30 dB

So far as the other comments are concerned, we answer these in the same sequence as the comments.

(*a*) For the BPSK modulations considered in our letter there is no quadrature phase component. Therefore, similar to Reference 3, we have used only the inphase component. For the quadrature amplitude modulation techniques both inphase and quadrature phase componets may be used.

(*b*) In our letter,[1] the interference is assumed to be a Gaussian random variable whose variance changes with the number of interferers. The integration is on all values of the interference random variable rather than the position of the interfering terminals. Our later work[2] includes assumptions similar to those suggested in the comment.

(*c*) The average error rate in the fading channels is dominated by the error rate in the deep fades. Our assumption, that in slow fading the bit error probability is the same as the packet error probability, is based on the observation that in the deep and slow fades the packets can not survive regardless of the length. This assumption is not supported by mathematical derivations. As far as the accuracy is concerned, our recent work[2] shows that the exact solution is close to this approximation.

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BALANCED 3 × 3 PORT PHASE DIVERSITY RECEIVER WITH REDUCED IMPACT OF THERMAL NOISE

Indexing terms: Optical communications, Optical receivers

The first 3 × 3 port balanced phase diversity receiver in which thermal noise has the same impact as in heterodyne receivers was realised although three preamplifiers are used. Experimental and theoretical results which prove the reduced impact of thermal noise compared with a conventional 3 × 3 port receiver are presented.

In high data rate and multichannel coherent optical communications, the phase diversity (PD) receiver is a promising candidate because of its ability for baseband reception.[1] It is essential that this receiver is balanced so that the local oscillator (LO) intensity noise (RIN) and other intensity fluctuations are suppressed. This is important in multichannel systems as all channels contribute to the latter disturances.[2]

We present experimental and theoretical results of the balanced ring structure 3 × 3 port PD receiver proposed in Reference 3 and realised for the first time. In this receiver thermal noise has the same impact as in heterodyne receivers although three preamplifiers are used. This balanced 3 × 3 port receiver offers a remarkably higher sensitivity over the conventional 3 × 3 port receiver. This advantage was not noticed in Reference 3 but has been proved in 565 Mbit/s DPSK transmission experiments.

The general structure of the balanced 3 × 3 port PD receiver is shown in Fig. 1. The optical output signals of the 3 × 3 fibre coupler are fed to three photodiodes arranged in a closed triangle. The photocurrents of neighbouring photodiodes are subtracted. The RIN of the LO is suppressed and the signal current amplitude increases by a factor of $\sqrt{3}$. The ratio of the thermal noise, generated in the following amplifier, to the signal power is reduced by a factor of three when compared with the conventional 3 × 3 port receiver. It is found that the shot noise contribution is almost the same in the balanced and conventional cases after the final summation of the three receiver branches. A receiver sensitivity gain of up to 4·8 dB is expected in the low LO power range where the thermal noise is dominant.

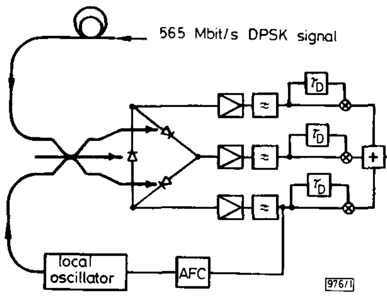

Fig. 1 *565 Mbit/s DPSK receiver with balanced 3 × 3 port front end*

For the conventional and the balanced receiver a theoretical analysis was performed assuming matched filters and white thermal noise. For the conventional receiver the results are already given in Reference 3. The result given in Reference 3 for the balanced receiver is correct only if all the noise is generated within the ring. Since the shot noise is generated at the photodiodes in the ring, and the thermal noise is generated at the preamplifiers outside the ring, both noise sources contribute to the output signal. Only analytic approximations can be derived. If the shot noise is dominant the BER is approximated by

$$BER = 0{\cdot}5 \exp(-SNR) \tag{1}$$

with $SNR = RP_S T/[e(1 + \overline{i_{th}^2}/2eRP_{LO})]$.

P_S is the signal power, P_{LO} is the LO power, $\overline{i_{th}^2}$ is the thermal receiver noise, R is the photodiode responsivity and e is the elementary charge. Eqn. 1 is equal to eqn. 5 of Reference 3 except for the reduction of the thermal noise by a factor of three.

If the thermal noise is dominant, the result is the same as for the conventional receiver with $RIN = 0$, but again with the thermal noise reduced by a factor of three. An asymptotic